TRANSMITTER DESIGN IN A COMMERCIAL MARINE SINGLE-SIDEBAND TRANSCEIVER.

by

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UNITED STATES
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THESIS

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ABSTRACT

Design of the transmitter portion of a solid state, state-of-the-art, single-sideband, 2 - 17 MHz transceiver is presented. A short discussion of the theory of single sideband and a comparison of single sideband systems with amplitude modulated systems are also presented. The unique requirements of commercial, marine communications are considered and the method of their fulfillment in this transceiver is discussed. Circuitry common to both the receiver and the transmitter is presented in detail.

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INTRODUCTION

Reliable, long-range communications depend primarily on ionospheric reflections of electromagnetic waves. The most efficient reflections occur in the high-frequency (h-f) band. Frequencies above the h-f band are usually well above the maximum useable frequency, hence they are normally restricted to line-of-sight transmission. Below the h-f band, ionospheric attenuation in the D and E layers becomes so great as to make these frequencies unuseable at ranges greater than can be obtained by ground wave propagation. This is true primarily during the hours of daylight since the D and E layers disappear at night. However, one generally wishes to communicate equally well during the daylight hours as during the hours of darkness, hence these frequencies below the h-f band are also unuseable. It should be noted here that very-low-frequency transmissions can provide very reliable long-range communications. However, high power and very large antennas are required in this frequency range. Also, bandwidth is so limited as to make voice communications most difficult if not impossible.

The h-f band, covering the range of frequencies from approximately 2 to 30 MHz, is therefore in great demand by amateurs, commercial interests, and the military. Thus any modulation scheme which requires a minimum of bandwidth, allowing a maximum number of channels, is a strong contender for use. Single sideband (SSB) is such a system. There are other advantages over the two other principal methods of modulation, i.e., amplitude modulation (AM) and frequency modulation

(FM). There are also disadvantages to SSB. A comparison of SSB with AM and FM, will show that the advantages outweigh the disadvantages, particularly with today's state-of-the-art technology. This comparison will be made in the next section.

The Federal Communications Commission (FCC), recognizing the advantages to be derived by the use of SSB, has ruled that after 1 January 1974, SSB radio telephone operation will be required in the maritime services, except those in Alaska, on frequencies between 4 and 27.5 MHz.[(1)] It is apparent then that any new equipment designed for mobile maritime use in this band should use SSB for voice communications. This thesis is concerned with the design and construction of the transmitter portion of a high-frequency, SSB transceiver which is to be submitted to the FCC for type acceptance.

Since the transceiver is to be designed for mobile, maritime use primarily aboard small fishing vessels, a great deal of consideration should be given to the size, weight, and power required for its operation. The desired output power level, number of channels, etc. have a bearing on these considerations but certainly solid-state devices can minimize them while also providing a more rugged construction. Accordingly the design objectives are that the transmitter be 12 channel, solid state (except for the driver and the power amplifier), capable of delivering 250 watts of peak-envelope power (PEP) to the antenna coupler, and that the entire transceiver occupy a space no greater than 18" wide, 8" high, and 10" deep. The complete list of specifications may be found in Appendix I.

The objective of designing, constructing, and testing the SSB transmitter portion of the transceiver was quite ambitious for the time available. Consequently certain portions of the transmitter were not fully developed, and the complete transceiver was not realized.

COMPARISON OF SSB WITH AM AND FM

To be meaningful, any comparison made between two or more systems must be on some common basis. In this section the comparisons will be made on the basis of required bandwidth, on the power required at the transmitter to produce an equal signal-to-noise ratio (SNR) at the output of the receiver under ideal propagation conditions, and on the distortion effects of non-ideal propagation conditions.

Wide-band FM may be dispensed with on the basis of required bandwidth since it is extremely wasteful of the available spectrum, and is not used in the h-f band. Narrow-band FM has essentially the same amplitude spectrum as AM, hence it may be included in the comparison of AM with SSB.

SSB is not a unique method of modulation, but is normally a form of AM. By current techniques it is derived from an AM, double-sideband, suppressed-carrier signal by suppressing one of the side bands. That a single side band of an amplitude-modulated wave is sufficient can be seen by expanding the equation for the AM wave. The expression for the carrier is

$$v_c = A_c \cos \omega_c t. \qquad (1)$$

To produce AM, one modulates the magnitude of the carrier with the information one wishes to transmit. Assume that the information signal to be transmitted consists of two tones, ω_1 and ω_2 radians, where ω_2 is greater than ω_1. Then

$$v_m = A_1 \cos(\omega_1 t) + A_2 \cos(\omega_2 t). \qquad (2)$$

If the carrier is amplitude modulated by the information signal, the resultant can be expressed by

$$v = A_c\,[1 + m_1\cos(\omega_1 t) + m_2\cos(\omega_2 t)]\cos(\omega_c t) \tag{3}$$

where the m's are the indices of modulation for each tone.

Expanding, one gets

$$\begin{aligned} v &= A_c\cos(\omega_c t) + m_1A_c\cos(\omega_1 t)\cos(\omega_c t) \\ &\quad + m_2A_c\cos(\omega_2 t)\cos(\omega_c t) \\ &= A_c\cos(\omega_c t) + (m_1A_c/2)\cos(\omega_c - \omega_1)t \\ &\quad + (m_2A_c/2)\cos(\omega_c - \omega_2)t \\ &\quad + (m_1A_c/2)\cos(\omega_c + \omega_1)t \\ &\quad + (m_2A_c/2)\cos(\omega_c + \omega_2)t. \end{aligned} \tag{4}$$

It can be seen from the above expression that all of the information about the amplitude and frequency of each component of the modulating signal is contained in either sideband provided the carrier frequency is known. Furthermore, it can also be seen that the bandwidth for AM must be twice the highest frequency of the modulating signal.

Since SSB requires only one sideband, the spectrum required is approximately half that required for AM. This is a very important consideration in the h-f band where there is a great demand on the available spectrum space.

SSB offers another big advantage over AM in that it requires significantly less transmitted power in order to achieve the same SNR at the output of the receiver. It has been found that under ideal propagation conditions, an SSB transmitter rated at 0.5 W of PEP will produce the same SNR at the output of a receiver as an AM transmitter rated at a carrier power of 1 W.[2]

Consider now the case of single-tone AM. Equation (4) then becomes

$$v = A_c\cos(\omega_c t) + (mA_c/2)\cos(\omega_c - \omega_1)t$$
$$+ (mA_c/2)\cos(\omega_c + \omega_1)t \qquad (5)$$

where the subscript on m has been dropped since it is no longer needed. The total average power of an AM signal is the sum of the average power of the carrier and its sidebands. Referring to equation (5), the total power is found by taking one half the sum of the squares of the magnitude of each frequency component and dividing by the resistance across which the voltage is developed. That is

$$P_t = [A_c^2 + m^2A_c^2/4 + m^2A_c^2/4]/2R. \qquad (6)$$

The first term on the right side of the equation is the carrier power. The second and third terms represent the power in each sideband. For 100% modulation, m = 1, and the power in each sideband is $A_c^2/8R$ W or one-fourth the power in the carrier. The total power for the AM signal is then 1.5 times the carrier power or three times the power of the SSB signal to achieve the same SNR at the receiver.

Coherent detection of the AM signal will produce a signal power in the receiver which is twice the power developed in the detection of the SSB signal. This apparent advantage for AM is offset, however, by the fact that the AM signal requires twice the bandwidth of the SSB signal, admitting twice the noise power. Thus the SNR remains the same.

A comparison should also be made of the peak voltages present at the antenna for the two systems, particularly for mobile equipment where the antennas are electrically short and corona breakdown may be a limiting factor. The peak voltage for a one-tone, 100% amplitude modulated signal is the sum of the peak voltages of the carrier and the two sidebands. From equation (5) one can see that the peak voltage is $2A_c$. The peak voltage required for the SSB signal to produce the same SNR at the output of the receiver is only $0.707A_c$ or about 1/3 the peak voltage of the AM signal.

SSB also has a clear advantage over AM in the presence of selective fading. Selective fading occurs when a signal arrives at the receiver via two or more paths whose lengths are unequal. In the case of two path propagation, the signals may arrive at the receiver either in phase, 180 degrees out of phase, or somewhere in between. If the path lengths are sufficiently different, which is frequently the case, the relative phases of the two signals are highly frequency sensitive, so that within the bandwidth of the AM or the SSB signal, some components are enhanced while others are degraded.

If the geometry of the situation is such that cancellation of the carrier occurs, the intelligibility of the AM signal becomes seriously degraded because the process normally used for detection requires a carrier. The SSB signal is unaffected by the loss of the carrier frequency since there is no carrier in the signal to begin with. Cancellation can occur equally as well at some other frequency in the signal. This loss does not seriously degrade either the AM signal or the SSB signal if the intelligence transmitted is voice.

In addition to cancellation of the carrier, AM signals are also degraded by a relative phase shift in the carrier. A 90 degree phase shift of the carrier results in the complete loss of the fundamental frequency of the detected signal when envelope detection is used, and distortion products alone exist. The distortion products are mainly even harmonics with the second harmonic being the most significant.

One can see that SSB offers many advantages over AM. There are, however, some disadvantages. The chief disadvantage is that detection of the SSB voice signal depends upon the reinsertion of the carrier at very nearly the correct frequency. The carrier frequency generated in the receiver must be within about ± 20 Hz of the original carrier frequency in the transmitter. Furthermore, some means must be provided for removing the carrier and one sideband. Both of these requirements serve to add to the complexity, hence the cost, of SSB equipment. It should be noted, however, that when cost is considered on the basis of dollars per watt of information signal, SSB becomes cheaper than AM at the higher power levels.

The major advantages of SSB are the narrow bandwidths, the lower power requirements, the lower peak antenna voltages, and reduced susceptability to distortion due to fading. The major disadvantages are complexity and high cost.

THEORY OF SSB GENERATION

This discussion of single-sideband generation will be oriented to the technique used in this transmitter. A block diagram of the transmitter is shown in Figure 1.

For purposes of this discussion, assume the audio waveform to be transmitted is as shown in Figure 2a. This signal is linearly amplified in the audio amplifier, and then fed into the balanced modulator along with the signal from the beat-frequency oscillator (BFO). The signal from the BFO is at the i-f frequency. The output of the balanced modulator is an AM double-sideband/suppressed-carrier signal as shown in Figure 2b. The position of the i-f carrier is indicated by the dashed spectral line. The i-f carrier is balanced out in the balanced modulator, hence its name. This signal is then fed into a narrow-band filter which passes one sideband but not the other. For this transmitter a lower sideband (LSB) filter is used since upper sideband (USB) transmission is desired and difference mixing is used for conversion to the final operating frequency. This causes the order of the frequency components to be reversed.

The LSB signal (shown in Figure 2c) is then linearly amplified in the i-f amplifier. This amplified signal is then fed into the final mixer along with the local oscillator (LO) signal. The LO is at the desired output frequency plus the i-f frequency. The mixer output, consisting of the LO signal, the i-f signal, the sum and difference of the LO signal and the i-f signal, and harmonics, are then fed through a series of tuned r-f amplifiers to the antenna. Since the sum and the difference frequencies are separated from the LO signal

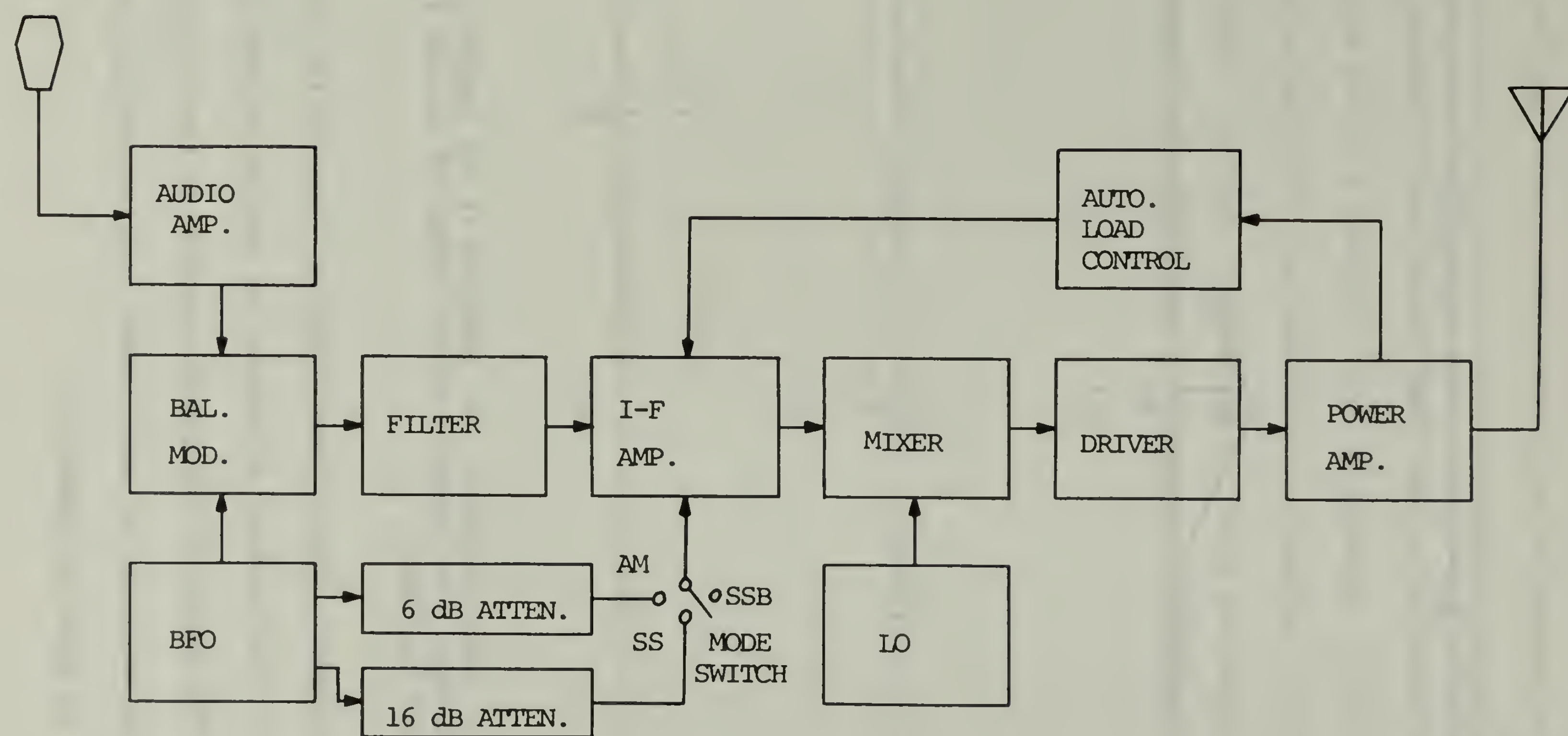


Figure 1. Transmitter block diagram.

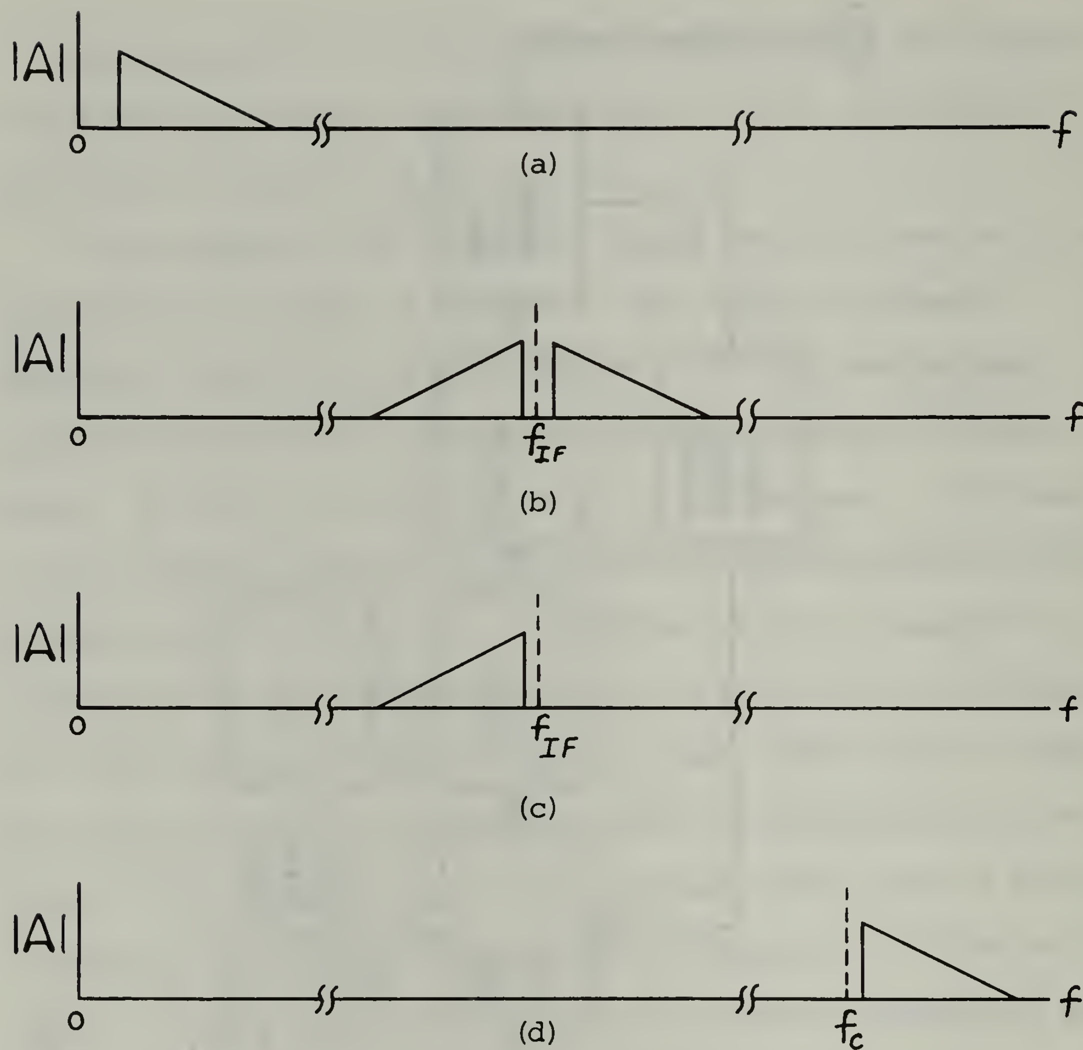


Figure 2. (a) Audio spectrum. (b) Spectrum at the output of the balanced modulator. (c) Spectrum at the output of the LSB filter. (d) Transmitted spectrum.

by the i-f frequency, high-Q tuned circuits, tuned to the difference frequency, will sufficiently attenuate the carrier and the sum frequency. This reinverts the signal spectrum so that all the frequency components are in their proper relative position for USB transmission as shown in Figure 2d.

Note that the basic SSB signal is generated at very low power levels and then linearly amplified up to the desired output power level after the final mixer. The i-f amplifier must also be linear.

The automatic-load-control (ALC) circuit performs much the same function as AGC in a receiver. ALC compensates for those operators who tend to shout into the microphone by reducing the gain of the i-f amplifier. It also compensates for those who whisper by increasing the gain. The net effect is to maintain a constant output power level.

TRANSMITTER DESIGN

GENERAL

The specifications governing the design of the transceiver are listed in Appendix I. This discussion will be confined to the transmitter section and that circuitry common to the receiver and necessary for the operation of the transmitter. For a detailed discussion of the receiver see reference 3.

The frequency range (2-17 MHz) of this transceiver is divided into three bands. The first two bands each cover an octave, and the third band covers a little more than an octave. The twelve discreet channels are divided into the three bands as follows:

Band 1 (2-4 MHz) - channels 1, 2, and 3,

Band 2 (4-8 MHz) - channels 4, 5, 6, 7, and 8,

Band 3 (8-17 MHz) - channels 9, 10, 11, and 12.

There is, however, no band switch on the front panel. The bands are automatically selected by the channel selector.

The twelve channels are further divided into two categories, i.e. simplex operation and semi-duplex operation. Simplex operation involves transmission and reception on the same frequency while semi-duplex operation involves transmission on one frequency and reception on another frequency, though not simultaneously. Channels 1, 2, 4, 5, 6, 9, and 10 employ simplex operation and channels 3, 7, 8, 11, and 12 employ semi-duplex operation.

Although the future of SSB in the h-f band is assured, there are still many transceivers in operation today which use AM transmission.

Thus new transceivers must be capable of generating and receiving AM. This transceiver does provide a compatible AM selection. Compatible AM is a SSB signal with a reinserted carrier which is 3 to 6 dB below the PEP of the sideband. This transceiver utilizes a carrier 6 dB below the PEP of the sideband.

In addition to the AM compatibility, there is also a requirement which is satisfied in this transceiver for carrier reinsertion 16 dB below the sideband for ship-to-shore transmissions.

The Rules and Regulations of the FCC require that output power be reduced to 150 W PEP in the 2-4 MHz range when engaged in ship-to-ship communications.(4) Furthermore, the compatible AM mode of operation is required on the emergency frequency of 2182 kHz. These requirements make either the operation of the transceiver or the internal circuitry fairly complex. Since this transceiver is to be used on board small fishing and pleasure vessels which probably would not have a professional radio operator aboard, the complexity is built into the transceiver by providing several automatic features which will be more fully discussed in the section on switching. Therefore the front panel has a minimum number of switches for operator simplicity. The front panel switches which are used are: (1) the volume control/ON-OFF switch, (2) the crystal ovens ON-OFF switch, (3) the filament power ON-OFF switch, (4) the channel selector, and (5) the mode selector. In addition there is a press-to-talk button on the microphone. Note that there is no "delta" frequency control; the frequency control circuits being precise enough to make this unnecessary.

The mode selector switch has three positions; AM for the compatible AM mode, SS for the ship-to-shore mode, and SB for the single-sideband mode.

AUDIO AMPLIFIER

The active device chosen for use as the audio amplifier is the RCA-CA3020 multi-purpose integrated-circuit wide-band power amplifier. Typical values of the characteristics which led to its being chosen are:

minimum zero-signal d-c current drain	14 mA
sensitivity	35 mV
power gain	58 dB
input resistance	40 kilohm
total harmonic distortion	3 %
signal-to-noise ratio	70 dB

The internal circuitry of the CA3020 may be represented by a functional block diagram as shown in Figure 3. The diagram shows that the CA3020 performs five functions; voltage regulator, buffer or optional amplifier, differential amplifier and phase splitter, driver, and power-output amplifier. Use of the buffer amplifier provides the high input

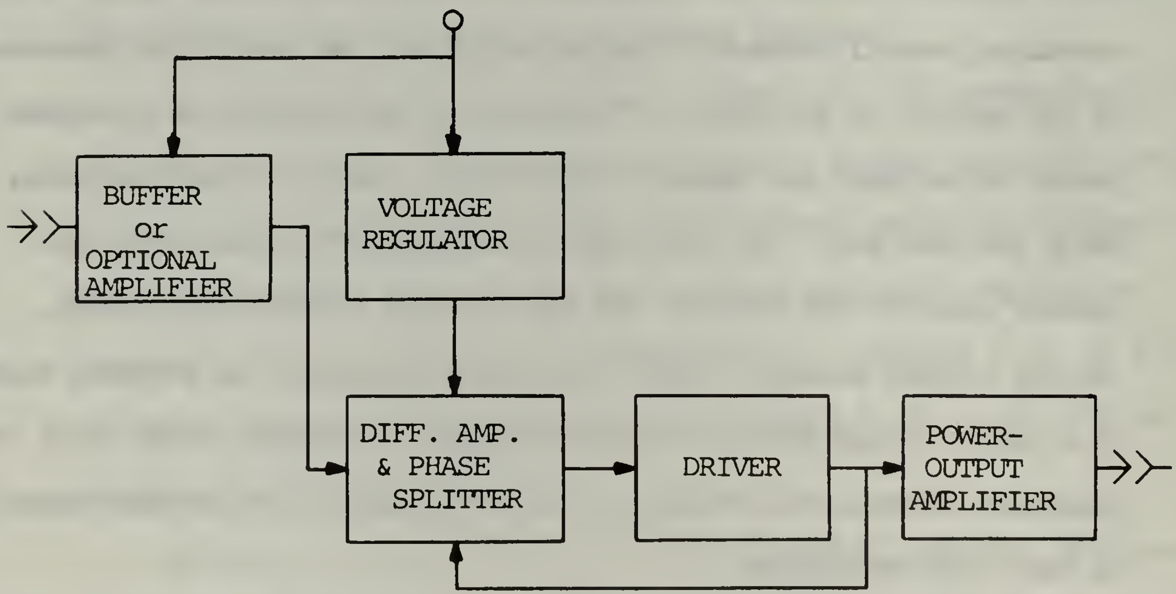


Figure 3. Block diagram of the CA3020 integrated circuit.

impedance to the device for the purpose of obtaining a better match with the high impedance microphone. A low input impedance of 600 ohms can be obtained by by-passing the buffer amplifier and feeding the signal directly into the differential amplifier. The differential amplifier amplifies the signal and effects the 180-degree phase shift necessary for the push-pull operation of the driver and power-output amplifier.

The driver consists of a pair of emitter-followers, one for each signal path. The feedback path is provided to assure that the d-c voltage between the collectors of the differential amplifier and phase splitter, is zero. Finally, the power-output amplifier delivers power to the load in a class B push-pull mode.

Temperature tracking is provided by the voltage regulator to maintain stable operation from -55 degrees Celsius to 125 degrees Celsius.

Figure 4 shows the CA3020 as it is used in the transmitter for amplification of the audio signal from the microphone. The bandwidth of the audio amplifier is restricted to the range of 300 Hz to 3 kHz by capacitors C_1, C_2, C_3, C_4, and C_5. Capacitors C_6 and C_7 are audio coupling capacitors while C_8 provides an audio by-pass to ensure that no audio signal gets into the d-c power supply.

R_1 is a biasing resistor for the buffer amplifier which is an emitter follower, with R_2 being the emitter resistor. R_2 provides an internal gain control adjustment over which the operator has no control. R_3 is used to help minimize distortion.(5) R_4 and R_5 are unby-passed resistors for the two transistors in the push-pull power-output amplifier

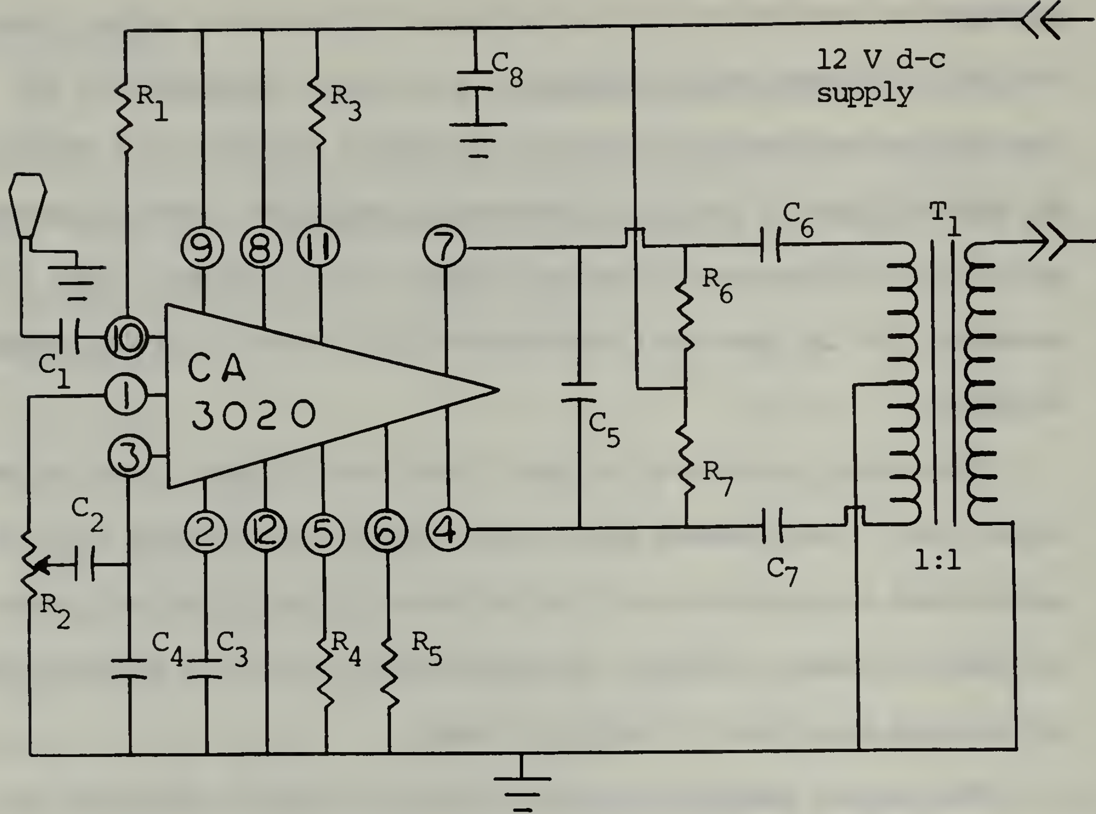


Figure 4. Audio amplifier.

stage of the CA3020. Introduction of this negative feedback also reduces distortion. Finally, R_6 and R_7 are the collector resistors of the push-pull power-output amplifier.

The audio transformer, T_1, couples the amplifier to the balanced modulator. Since the power-output amplifier of the CA3020 draws a d-c current far in excess of the 3 mA maximum rating of the transformer, d-c isolation is required.

The harmonic distortion of the audio amplifier was measured at a frequency of 300 Hz with a peak output signal of 2.0 V. First the harmonic components of the signal generator were measured. Then the transformer was coupled directly to the signal generator and the harmonic content of its output measured. Finally the amplifier was connected as shown in Figure 4 and the harmonic content of the transformer output again measured. The results of these measurements are tabulated below with the level of the fundamental frequency as the reference.

Equipment used:

Audio Signal Generator - HP model 200AB ser. #310-18233

Oscilloscope - Tektronix type 515A ser. #3150

Wave Analyzer - HP model 302A ser. #724-05716

Results:

Harmonic	Source	Transformer	Amplifier
2 nd	-54.0 dB	-54.0 dB	-40.0 dB
3 rd	-62.5 dB	-62.5 dB	-49.0 dB
4 th			-71.0 dB
5 th			-66.0 dB
6 th			< -72.0 dB

Analysis of the above results shows that the actual harmonic distortion within the amplifier itself is -40.0 $\pm$ 0.2 dB for the second harmonic and -49.0 $\pm$ 0.2 dB for the third harmonic. The limits of $\pm$ 0.2 dB arise from the fact that it is not known if the distortion in the amplifier is in the same phase as the distortion in the generator. At any rate the distortion is well within acceptable limits.

BALANCED MODULATOR and FILTER

The balanced modulator, shown in Figure 5, is of the ring diode type using germanium diodes of the 1N34A type.

Balance is accomplished by two controls, R_9 and C_{11}. R_9 adjusts the magnitude while C_{11} adjusts the phase so that a more complete balance may be obtained.

To understand the operation of the balanced modulator of Figure 5, one must realize that C_9 must be in series resonance with the top half of the primary winding of T_2 at the i-f frequency. Likewise, the parallel combination of C_{10} and C_{11} must be in series resonance, or very nearly so, with the bottom half of the primary of T_2 at the i-f frequency. Series resonance is required in order to present a very low impedance path. The series resonant path is in parallel with a forward biased diode, which, in itself, is a low impedance path. It is also important to realize that C_{12} must be of such a value as to be a virtual short circuit to the i-f frequency and yet be an open circuit to the audio signal.

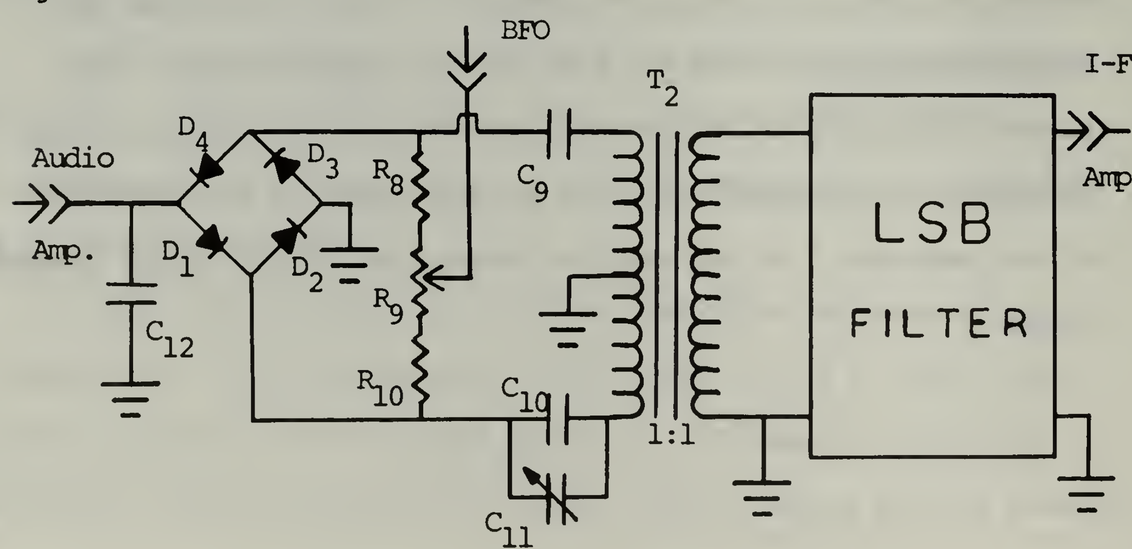
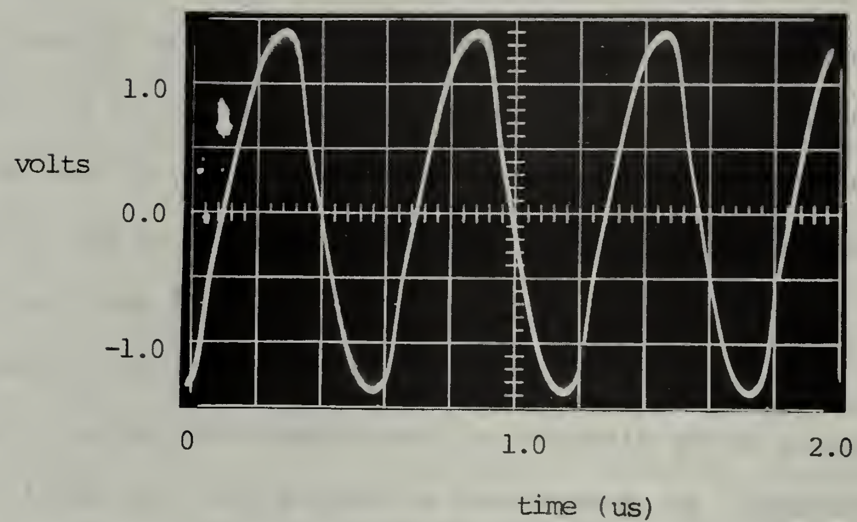


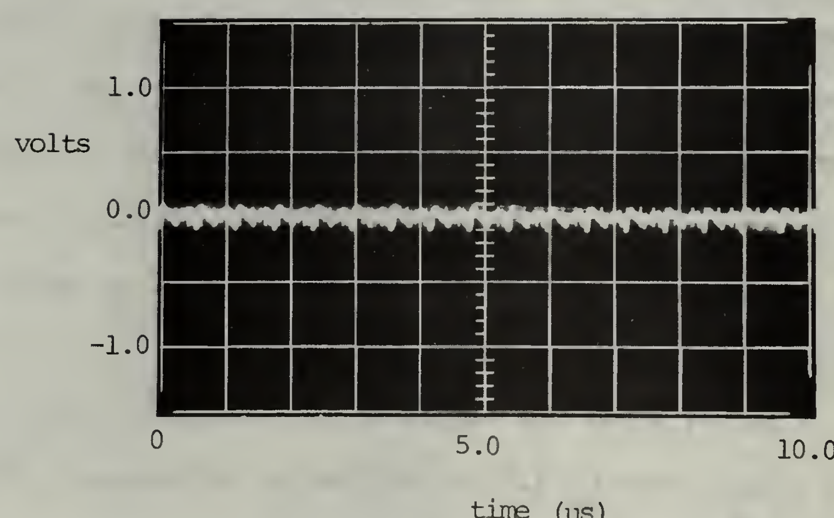
Figure 5. Balanced modulator.

The i-f carrier is fed into the balanced modulator through R_9 and divides into two paths, one through R_8 and the other through R_{10}. When the polarity of the BFO signal is positive, diodes D_2 and D_4 conduct, connecting the audio to the upper half of T_2, while diodes D_1 and D_3 are switched off. When the polarity of the BFO is negative, diodes D_1 and D_3 conduct connecting the audio to the lower half of T_2, while diodes D_2 and D_4 are off. Note that when diodes D_1 and D_3 are conducting the direction of the current in the primary of T_2 is opposite to the direction of the current when diodes D_2 and D_4 are conducting. It is important to realize that the magnitude of the BFO signal must be greater than that of the audio signal so that it is the carrier, and not the audio, which controls the switching of the diodes. Since the audio voltage is impressed across the primary of T_2 with one polarity when D_2 and D_4 conduct and with the opposite polarity when D_1 and D_3 conduct, the peak-to-peak voltage developed across the primary of T_2 is twice the peak voltage of the audio signal. Therefore, if the magnitude of the audio signal is zero, there should be no output from the balanced modulator. Figure 6b shows the output of the secondary of T_2 under the condition of no audio input. The injected BFO signal is shown in Figure 6a.

Figure 7 shows the output at the secondary of T_2 with about a 1.3 V audio tone of 2150 Hz applied to the balanced modulator. Note that the output of the balanced modulator is a two tone signal, one tone for each sideband. This signal then passes through the LSB filter which removes the USB. The output is then a single tone for a single audio tone input to the balanced modulator.



(a)



(b)

Figure 6. (a) BFO input. (b) Output of the balanced modulator with no audio input.

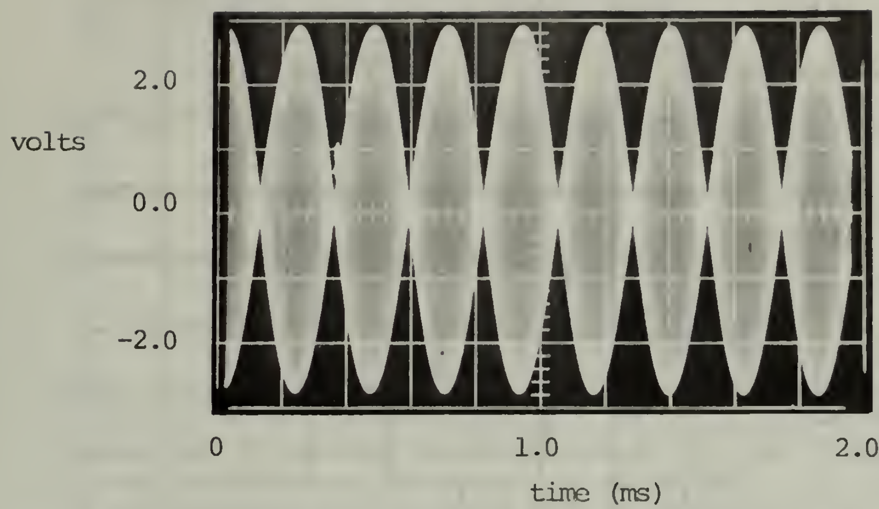


Figure 7. Balanced modulator output with a single audio tone input of 2150 Hz.

The filter used in this transmitter is a crystal filter manufactured by Filtech Corporation. Figure 8 presents a graph showing the pass band characteristic of the filter. The principal characteristics are listed below:

Carrier frequency:	1650 kHz
3 dB bandwidth:	2.1 kHz nominal
Carrier frequency rejection:	25 dB minimum
50 dB bandwidth:	4.2 dB maximum
Ripple:	2 dB
Input impedance:	500 ohms
Output impedance:	500 ohms
Insertion loss:	5 dB
Operating temperature:	-10° C to 80° C.

The output of the filter is then capacitively coupled to the i-f amplifier which is discussed in the next section.

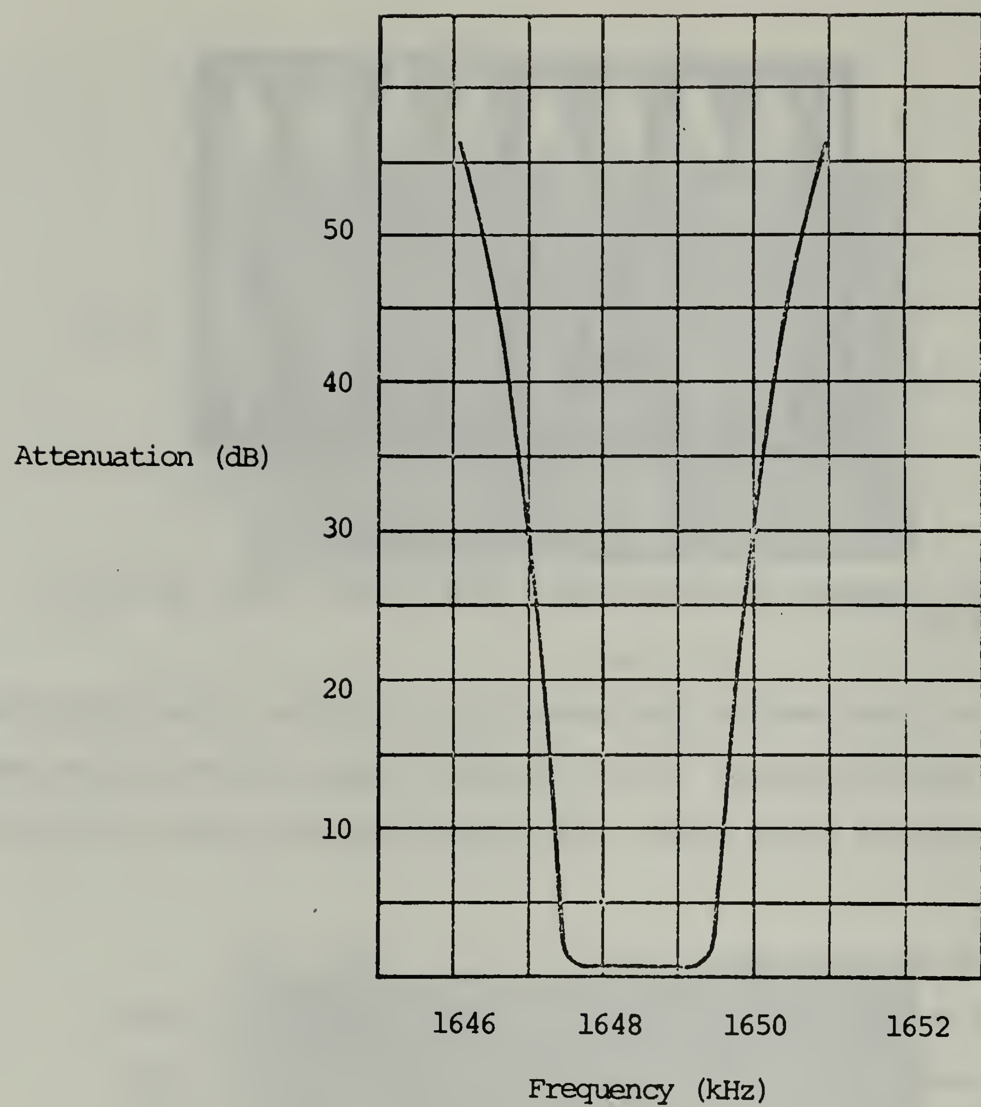


Figure 8. Characteristics of the Filtech LSB filter. Center frequency 1650 kHz.

I-F AMPLIFIER

RCA provides the integrated circuit used as the i-f amplifier for this transmitter. The device chosen is the CA3002 which is also used as the i-f amplifier in the receiver. The internal circuitry is presented in block diagram form in Figure 9.

Input signals may be applied to the CA3002 in either a push-pull mode, terminals 5 and 10, or a single-ended mode at terminal 10. Both inputs are to emitter followers which provide a high input impedance of approximately 100 kilohms.[6] Amplification is provided by the balanced differential amplifier which is fed from a constant current source. The constant current source provides an excellent means for controlling the gain of the amplifier by controlling the bias of the differential amplifier. This is accomplished by controlling the voltage between terminals 1 and 2. The output is single-ended from another emitter follower, providing an output impedance of approximately 80 ohms.[6] Either a direct coupled or a capacitively coupled output is provided.

The CA3002 may be operated at ambient temperatures ranging from -55°C to 125°C. It has a built in temperature compensating network for stabilization of the d-c operating point and gain over this temperature range.

Operation of the CA3002 may be from either single or dual power supplies. The dual supply, providing both a positive and a negative voltage, need not be symmetrical. For any bias condition, voltage between terminals 1 and 2, there are four modes of operation possible. Each mode has a distinct d-c operating point with a characteristic temperature

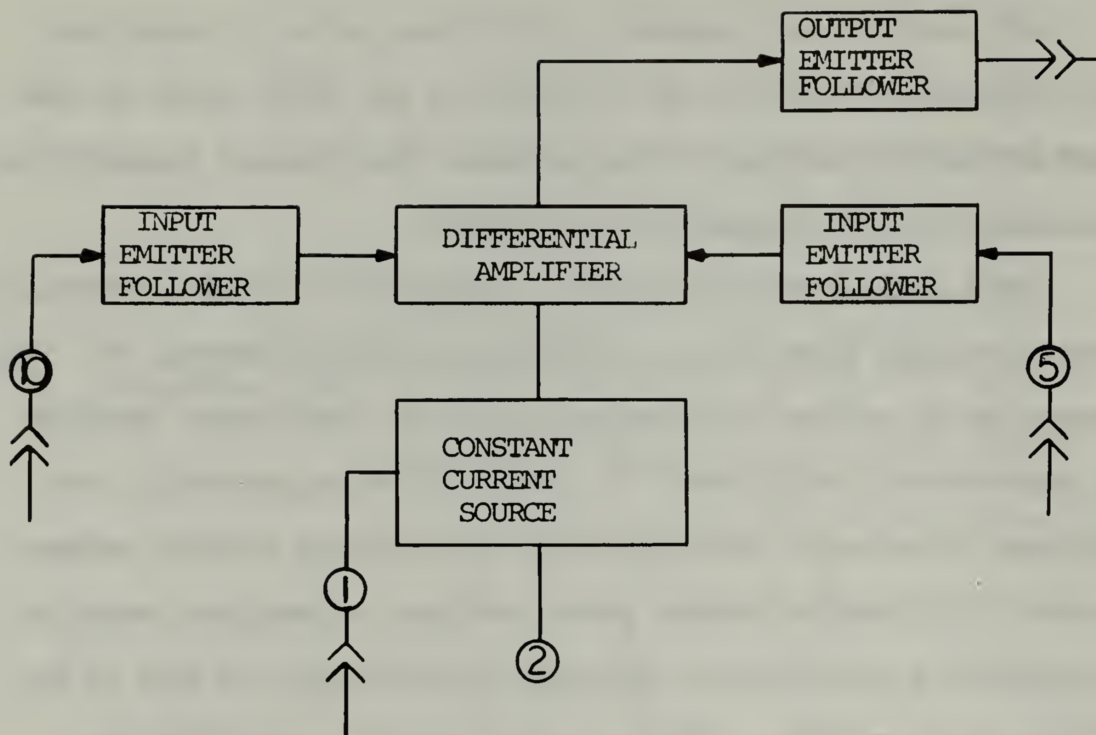


Figure 9. Block diagram of the CA3002 integrated circuit.

dependence and a distinct value of gain with its characteristic temperature dependence. The d-c operating point refered to is at the output terminal of the device and is of most concern if the output is direct coupled to the next stage. Since the i-f amplifier, as used in this transmitter, utilizes a capacitively coupled output, curves showing the temperature dependence of the d-c operating point are of little concern and are not presented. Figure 10, however, shows the temperature dependence of the gain for the mode of operation employed. It can be seen that the variation of gain over the full temperature range is slight.

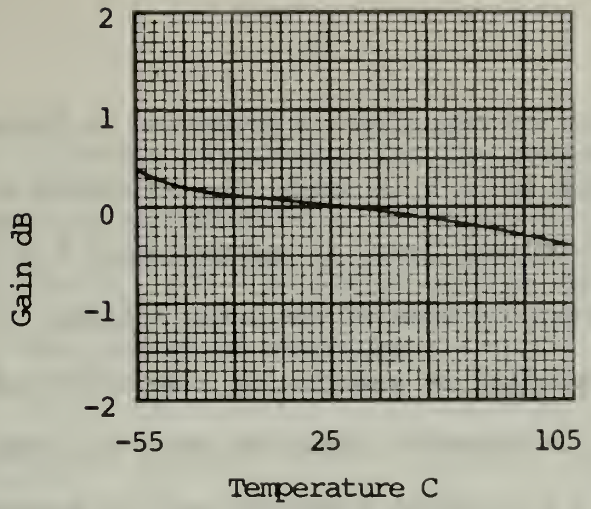


Figure 10. Gain of the CA3002 as a function of temperature. The reference level is 26 dB.

As previously indicated, control of the voltage applied to terminal 1 provides excellent gain control of the CA3002. This can be utilized to provide automatic load control (ALC) for the transmitter and to provide for the automatic reduction of transmitted power as required by the specifications. The switching arrangement necessary for this is presented in a later section. Figure 11 shows the AGC, or in this case the ALC,characteristics of this device at a frequency of 1.75 MHz. The reference level of gain is 26 dB. The ALC range is frequency dependent, varying from 75 dB at 1 MHz to 60 dB at 25 MHz.[6]

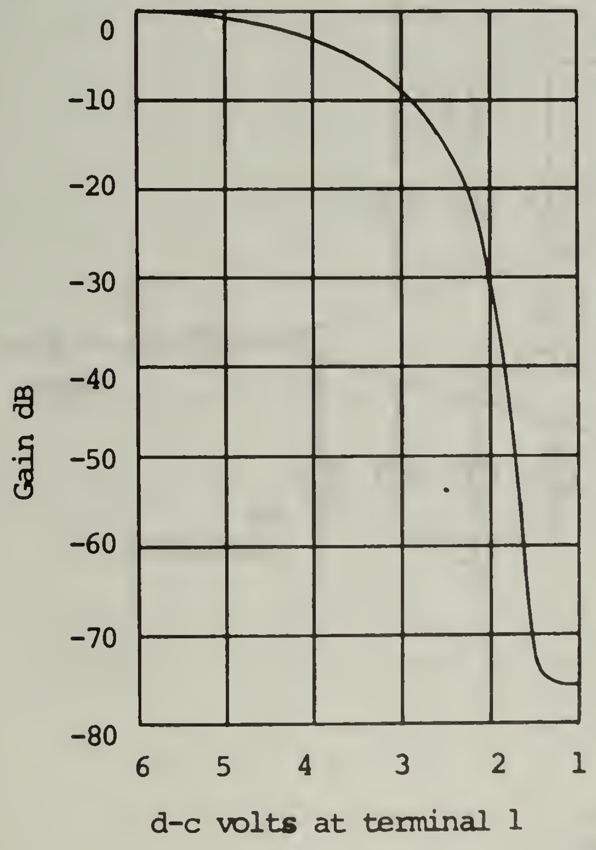


Figure 11. Gain of the CA3002 as a function of the d-c voltage at terminal 1 (normalized to a gain of 26 dB).

A schematic diagram of the i-f amplifier is presented in Figure 12. A single d-c supply voltage of 12 V is used. R_{11} in series with Z_1 provides the other required voltage level of approximately 6.1 V. R_{12} and R_{13} provide the necessary d-c bias of the input emitter followers. The gain of the amplifier is essentially independent of the values of R_{12} and R_{13}, but is dependent upon the ratio of these two resistors.(6) A ratio of 1.0 provided a maximum gain of approximately 15 dB with an ALC level of 3 V applied to terminal 1. Capacitors C_{13} and C_{15} are coupling capacitors from the crystal filter and to the mixer, respectively, while C_{14} is a by-pass capacitor to keep the i-f signal isolated from the power supply.

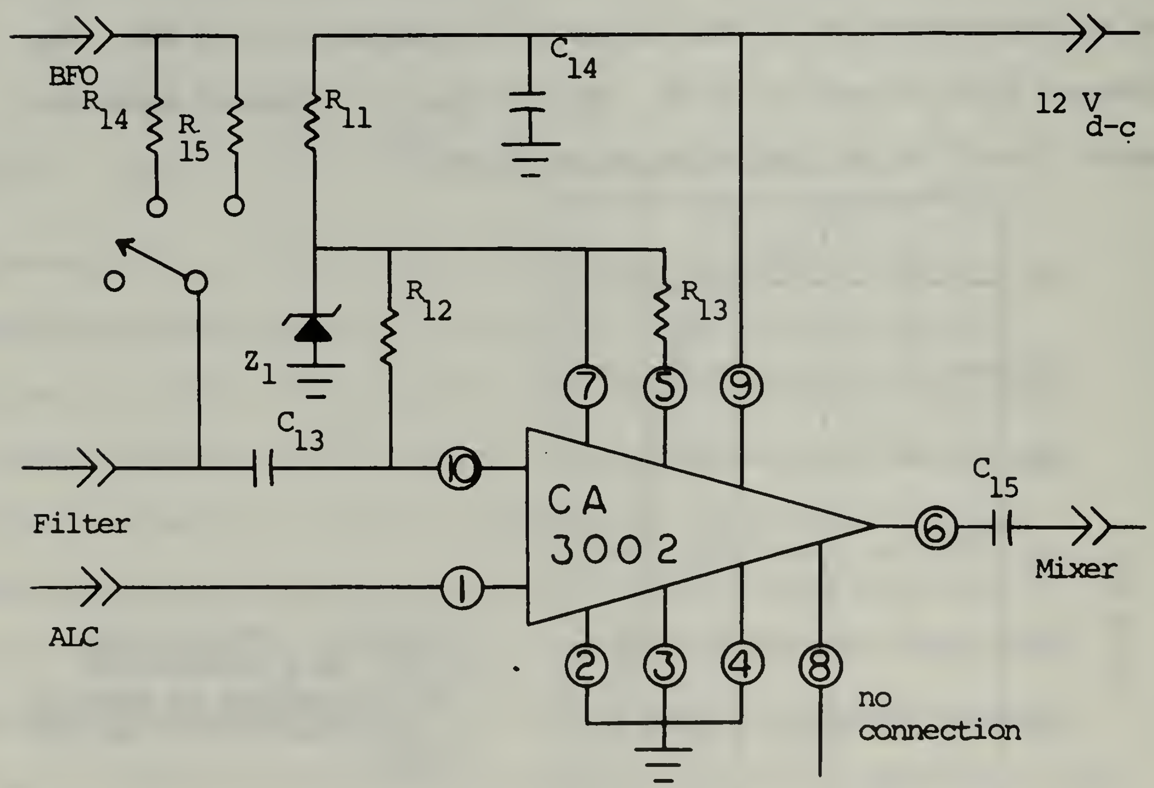


Figure 12. I-F amplifier.

Carrier reinsertion for the compatible AM and ship-to-shore modes is accomplished at the input to the i-f amplifier through R_{14} and R_{15} and a switching network which is discussed more fully in the section on switching.

MIXER

The final mixer is used to translate the i-f signal up to the desired output frequency. Mixing produces frequency components in the output which include the applied signals, sum and difference frequencies, and harmonics. In general one uses the difference between the local oscillator frequency and the i-f, particularly when a wide range of output frequencies must be accommodated. This keeps the relative frequency range of the local oscillator to a minimum.

As an example, consider the frequencies applicable to this transmitter. The output frequencies range from 2 to 17 MHz and the i-f is 1.65 MHz. Sum mixing would require the local oscillator to be capable of oscillating at a low frequency of 350 kHz and a high frequency of 15.35 MHz, or at a high frequency 38.4 times the low frequency. In contrast, difference mixing requires the local oscillator to operate from 3.65 MHz to 18.65 MHz, a ratio of only 5.1. An even smaller ratio could be obtained by using difference mixing at low frequencies and sum mixing at high frequencies. This technique, however, would require an LSB filter at those frequencies where difference mixing is used and an USB filter for those frequencies where sum mixing is used. The added complexity and cost would more than offset the slight advantage gained. Therefore difference mixing is used in this transmitter.

Since there are so many unwanted frequencies generated in the mixing process, tuned circuits must be provided to select the desired signal. It is here, at the output of the mixer, that the first distinction of frequency bands is made. The mixer of Figure 13 shows just one tuned circuit for simplicity. In actuality there are three coils, one for each band, and twelve capacitors, one for each channel. The appropriate coil and capacitor are switched into the circuit by the channel selector switch as described in the section on switching.

The mixer of Figure 13 utilizes a transistor, type 2N3118, as the non-linear element which is necessary for mixing to take place. In effect the transistor is a diode converter followed by an amplifier. There are three desirable characteristics which a transistor used as a mixer should have. These are:

1) efficient base-emitter diode characteristics,

2) low input capacitance, and

3) good gain at the desired signal frequency.[7]

The 2N3118 has these desirable characteristics.

The local oscillator signal, usually several times as large as the i-f signal, modulates the non-linear impedance of the base-emitter junction which forms the diode converter. This produces the mixing action when the i-f signal is applied. The transistor then amplifies the signals present. Note, however, that it is not important to provide amplification at all frequencies, but only at the desired output frequency. Thus a transistor may be used as a mixer even though it is not capable of amplifying either the i-f signal or the local oscillator signal, so long as it will amplify the desired output signal.

Referring to the mixer of Figure 13, inductors L_2 and L_3 are radio frequency chokes (RFC). They are used to allow the LO signal to be developed across L_2 and the i-f signal to be developed across L_3, while setting both the base and the emitter of Q_1 at d-c ground potential. When the LO signal goes positive it drives the base-emitter diode over the most non-linear portion of its characteristic curve, causing mixing to take place. L_1 is also an RFC across which the output signal is developed.

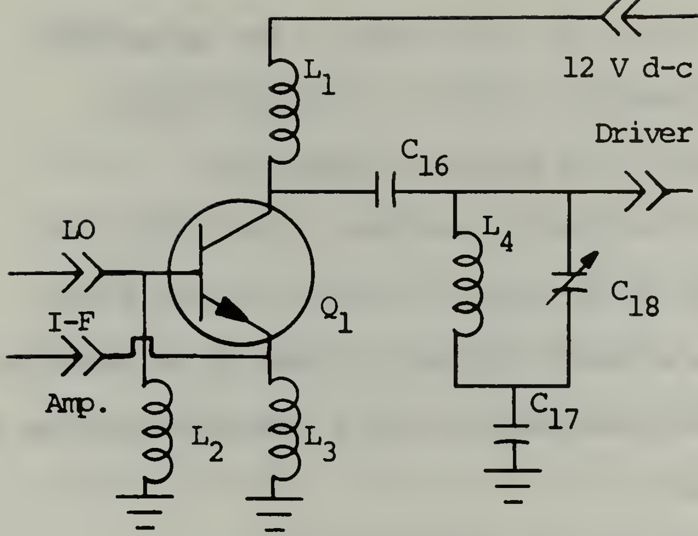


Figure 13. Final mixer.

The output is then coupled through C_{16} and a switching network (not shown) to the appropriate tuned circuit for the channel desired. Capacitor C_{17} is used for neutralization of the next stage.

Figure 14 shows the output waveform of the mixer with a two-tone audio input to the audio amplifier. The i-f level at the input to the mixer is 0.1 V peak, while the LO injection level is 1.0 V peak at a frequency of 9.5 MHz. The desired output frequency is 7.85 MHz. The halo effect is due to the relatively strong LO signal which is still present at this point. Further filtering removes this component. The spectrum of the desired signal at the mixer output is shown in Figure 15. It can be seen that the carrier level is approximately 45 dB below the sideband tones and the strongest intermodulation component at f_1 is about 38 dB below the desired sideband tones. This equates to the carrier being about 51 dB below the PEP of the sideband and f_1 being about 44 dB below the PEP of the sideband.

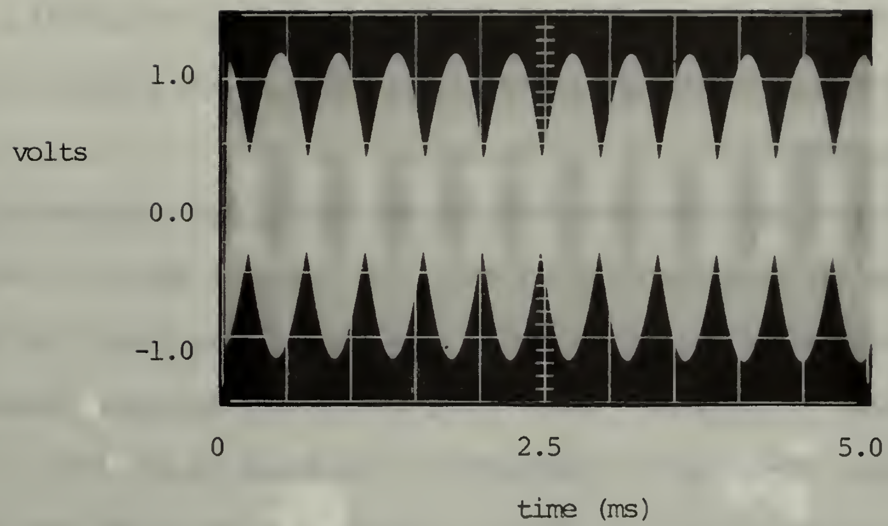


Figure 14. Waveform at the output of the final mixer.

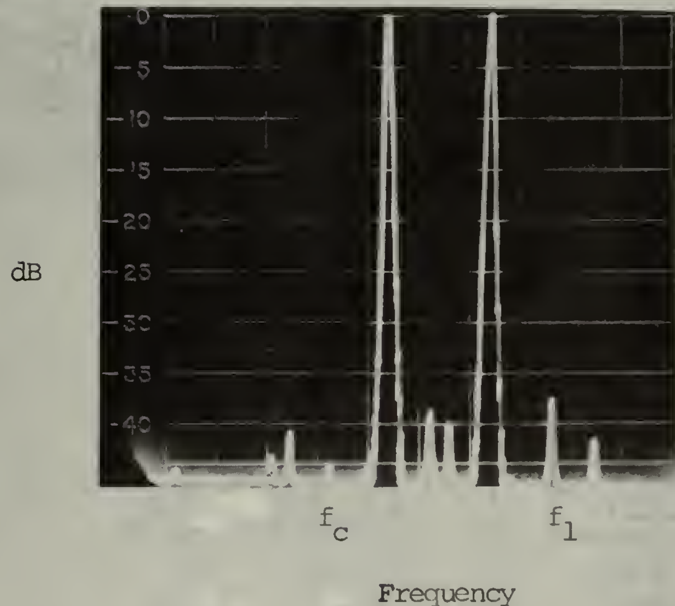


Figure 15. SSB spectrum at the output of the final mixer. Sweep width is 7 kHz. f_c = 1650 kHz.

DRIVER

After the desired SSB signal has been generated at low power levels in the exciter, it must be linearly amplified to a level suitable for driving the power amplifier. This is the function of the driver stage.

The driver amplifier was designed and built, but unfortunately it had a strong tendency to oscillate. The cause of the oscillations was not ascertained in the available time through an attempt was made to neutralize the circuit. This attempt to neutralize was, however, unsuccessful. Therefore no performance data could be obtained.

The proposed circuit is shown in Figure 16. Note that the tuned circuit at the grid of the driver is the same as was shown at the output of the mixer in Figure 13. The output of the driver is coupled

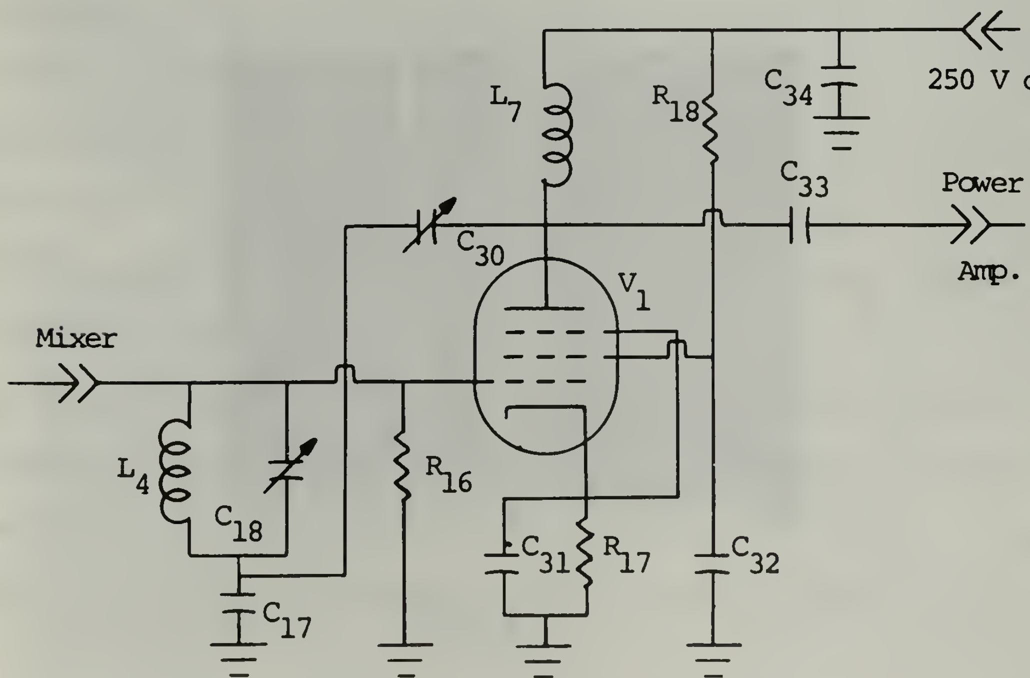


Figure 16. Driver amplifier.

through C_{33} to the primary of T_3 (shown in Figure 17), the secondary of which is tuned. Shunt feed of the d-c supply through L_7, an RFC, is employed in order to keep the d-c current out of the primary of T_3 to prevent the toroidal core from operating on the non-linear portion of its B-H curve. Resistors R_{16}, R_{17}, and R_{18} bias the tube, V_1, at the following d-c levels:

Plate voltage	250 V
Screen grid voltage	150 V
Control grid voltage	-10 V
Plate current	25 mA
Screen grid current	6 mA.

C_{31}, C_{32}, and C_{34} are r-f by-pass capacitors. C_{30} is a neutralizing capacitor. This same scheme for neutralization was used successfully in the power amplifier and will be fully explained in that section.

V_1 is a 12BY7A, a miniature, sharp cut-off pentode with a fairly high gain. This circuit, once neutralized, is capable of providing enough signal voltage at the grid of the power amplifier to develop the required output power.

POWER AMPLIFIER

The power amplifier is the final linear amplifier which delivers the required power to the antenna. Figure 17 shows the power amplifier developed for this transmitter.

The active component of this amplifier is a pair of 6LQ6 tubes driven in parallel in a class B mode. These are high-perveance beam-power tubes designed for use as the horizontal-deflection amplifier in color TV receivers. They are capable of withstanding a plate dissipation of 200 watts each for a period of 40 seconds, though their normal rating is only 30 watts of continuous plate dissipation.

For the two tubes in parallel the total rated continuous plate dissipation is 60 W. Considering a 50% plate efficiency for a two-tone signal (the theoretical maximum efficiency for class B operation with a two-tone signal is 61.7%),[(9)] the tubes could indefinitely deliver an average power of 60 W. This seems a far cry from the required 250 W PEP. However, for a two-tone signal the average power is 1/2 the PEP or 125 W. For a 50% transmit-receive duty cycle the average power required drops to 62.5 W. If one considers an additional duty factor to account for momentary pauses while actually transmitting (the case for voice signals), then the average power required falls within the rated capabilities of the tubes. This is because the power dissipated by the tubes is very low with no signal applied to the control grids.

For a voice signal the plate efficiency of the tubes will be much less but then so will the average output power required. The power dissipated by the tubes will also be less because of the mode of operation employed. The peak power handling capability of the tubes should be adequate to handle peak loads with ease, particularly when one considers that voice peaks are relatively rare.

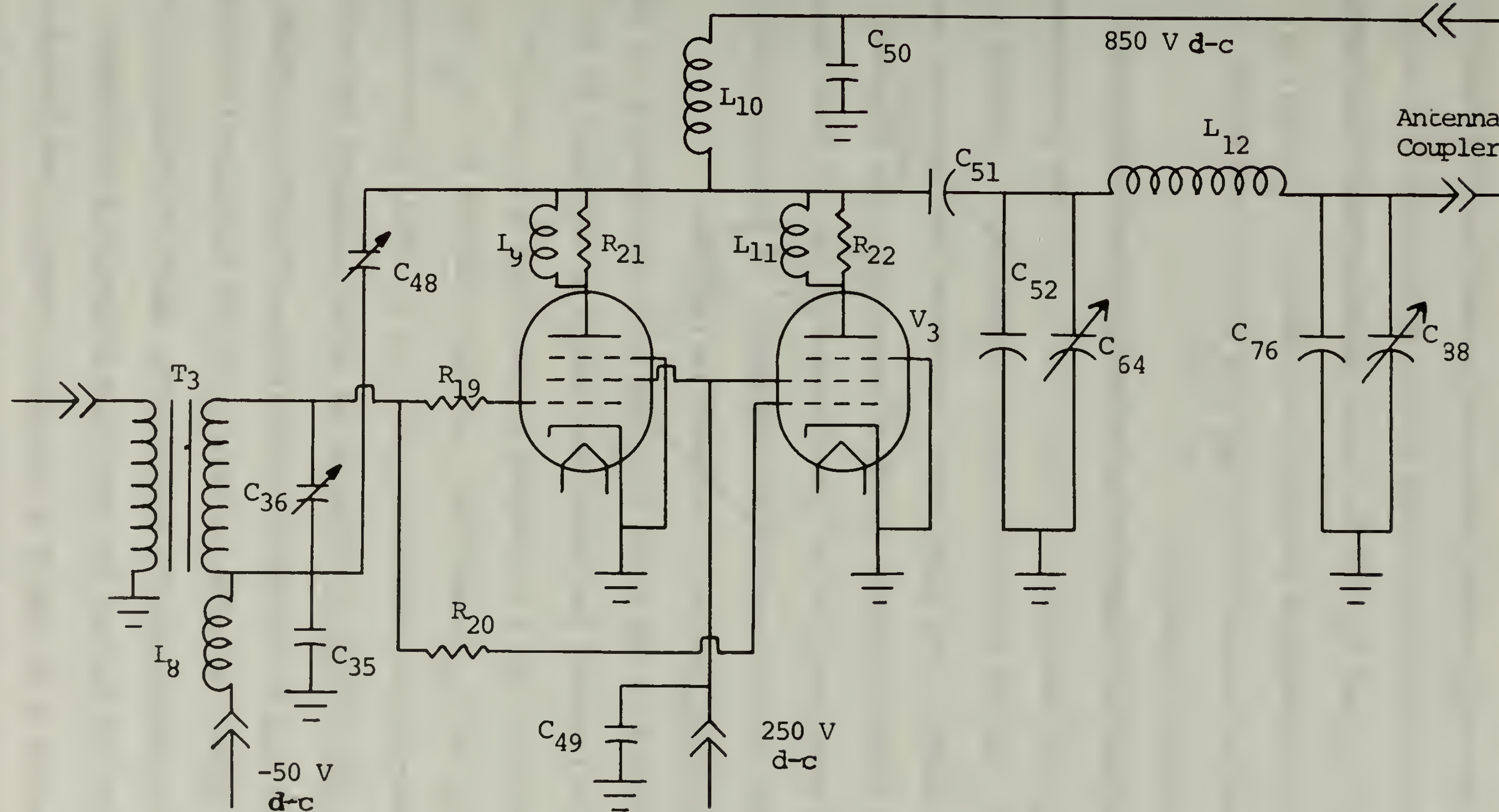


Figure 17. Power amplifier.

The 6 V heaters of the two tubes are connected in series to the 12 V d-c supply. Other required d-c voltages are (1) 850 V plate supply, (2) 250 V screen grid supply, and (3) -50 V grid bias supply.

R_{19}, R_{20}, R_{21}, and R_{22} are low value resistors used in conjunction with L_9, and L_{11} to suppress parasitic oscillations. C_{49}, C_{50}, L_8, and L_{10} are used to isolate the d-c supplies from the r-f signal. C_{51} is a d-c blocking capacitor.

Figure 17 shows one coupling transformer, T_3, and one tuning capacitor, C_{36}, for simplicity. There are actually three transformers, T_3, T_4, and T_5, one for each band, and 12 capacitors, C_{36} through C_{47}, one for each channel. The appropriate transformer and capacitor are chosen by the channel selector switch as described in the switching section. The turns ratio of the transformers has not yet been determined since the driver stage was not completed. It is anticipated that at most a 2 to 1 step-up ratio would be required.

Neutralization of the power amplifier is accomplished by C_{35} and C_{48}. Figure 18 presents the equivalent circuit at the input of the tubes. Points P, G, and X correspond to the plate, grid, and the junction of C_{35} and C_{48} respectively, as seen in Figure 17. C_{pg} is the total plate-to-grid capacitance and C_g is the total grid-to-cathode capacitance. The cathode is connected to ground. Without the neutralizing capacitor, C_{48}, in the circuit and with C_{35} replaced by a short circuit, plate signal current can pass through C_{pg} to the input tuned circuit at the grid of the tubes, causing regeneration. With the circuit of Figure 18 however, one can balance the capacitive bridge by appropriately choosing C_{35} and C_{48} so that no plate signal current flows through the input tuned circuit. For a given tube, or in this case pair of tubes in parallel, the values of C_{pg} and C_g are known. The criteria for choosing the appropriate values of C_{35} and C_{48} is that the ratio of these two capacitances must be equal to the ratio of the corresponding tube capacitances. That is,

$$\frac{C_{48}}{C_{35}} = \frac{C_{pg}}{C_g} .$$

This method of neutralization proved very effective for the power amplifier.

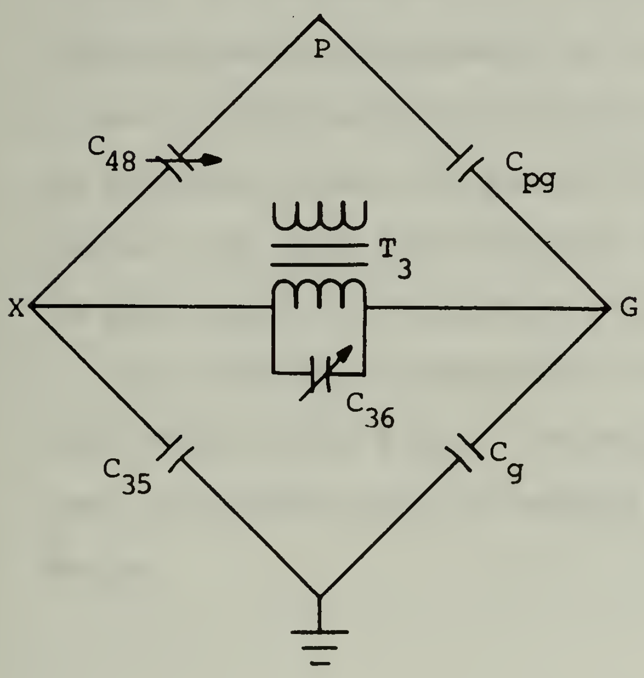


Figure 18. Equivalent neutralization circuit.

The pi network shown in Figure 17 is used to suppress undesired harmonics of the r-f signal and for impedance transformation. The two 6LQ6's in parallel require a load impedance of 900 ohms which is transformed to 50 ohms at the output of the pi network. There is at this point a switching network which is not shown in Figure 17 for simplicity, but which is detailed in Figure 23 in the section on switching. A detailed description of the design of the pi network is included as an appendix.

Preliminary testing showed that the power amplifier is indeed capable of delivering the required 250 W PEP to a 50 ohm resistive load at the output of the pi network with little or no inter-modulation distortion. However the antenna coupler would have to provide additional harmonic suppression in order to meet the specifications of the Federal Communications Commission.

AUTOMATIC LOAD CONTROL (ALC)

ALC is needed to maintain the peak-power output at a constant level as discussed in an earlier section. Due to lack of time no detailed design was developed for this circuit.

The general plan of attack, however, was to pick off a sample of the transmitted signal through a capacitive divider at the input to the pi network. This signal would then be detected in a standard diode detector and a d-c voltage proportional to the output power developed. This d-c voltage would then be compared with a threshold voltage. If the detected d-c voltage exceeded the threshold level, indicating excess output power, the voltage applied to terminal 1 of the i-f amplifier would be reduced by an amount proportional to the difference between the detected d-c voltage and the threshold voltage. This would reduce the gain of the i-f amplifier, and hence the power output would be reduced to the desired level.

The threshold level would have to be determined for each of the three desired output powers. The devices used to set these threshold levels would then have to be switched in and out of the circuit as desired.

OSCILLATORS

There are three oscillators used in this transceiver. One is the beat-frequency oscillator (BFO) and the other two are the local oscillators (LO). Two local oscillators are required because of the requirement for semi-duplex operation. All are Pierce crystal oscillator circuits utilizing metal-oxide field-effect transistors (MOSFET).

MOSFET's are used as the active device in the oscillator circuits because of their faster and better thermal stability which is of prime concern where oscillators are switched on and off as is the case with semi-duplex operation. The actual device used is the RCA 40468, an N-channel, depletion type MOSFET.

The BFO and the LO derive their names from their application to the receiver. That is, in the receiver the BFO is used as a BFO and the LO is used as an LO. In the transmitter the BFO signal is injected into the balanced modulator and the LO signal is used in the mixer for conversion of the i-f signal to the final output frequency.

BEAT FREQUENCY OSCILLATOR

Figure 19 shows the schematic diagram of the BFO. The FET is biased by the diode D_5 with a gate-leak resistor, R_{23}. This method of bias was used because of its superior output-voltage regulation. Since the frequency of oscillation of the BFO is below 2 MHz, namely 1.65 MHz, a capacitive divider across the crystal, X_1, is required.[8] The connection between the voltage divider capacitors, C_{100} and C_{101} in series with C_{102}, must be grounded. The additional capacitive voltage divider, C_{101} and C_{102}, from the drain of Q_2 to ground, is

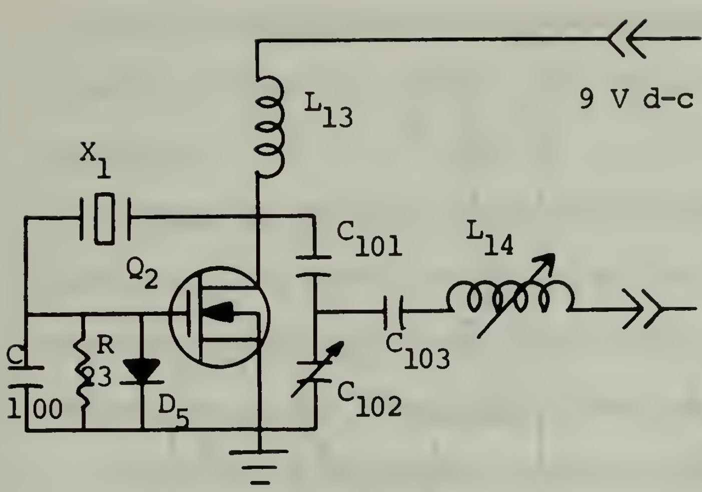


Figure 19. Beat-frequency oscillator.

provided to pick off a reduced output voltage. The desired output voltage is set at about 3 V peak-to-peak by adjusting C_{102}. C_{103} and L_{14} form a series resonant circuit which is used to suppress the harmonics of the beat-frequency oscillation. It can also be used to aid in setting the output level by slightly detuning L_{14}. L_{13} is an RFC which acts as a load for Q_2 and serves to keep the oscillations out of the d-c power supply.

The output of the BFO is always connected to both the transmitter and the receiver so that the load remains constant. However, the BFO is turned off by disconnecting the d-c power supply when receiving in the AM mode.

LOCAL OSCILLATORS

Two local oscillators are required because of the requirement for semi-duplex operation, transmission on one frequency and reception on another frequency. It should be mentioned that semi-duplex operation can be accomplished with one oscillator, but this would involve a much more complicated switching arrangement with much of the switching done

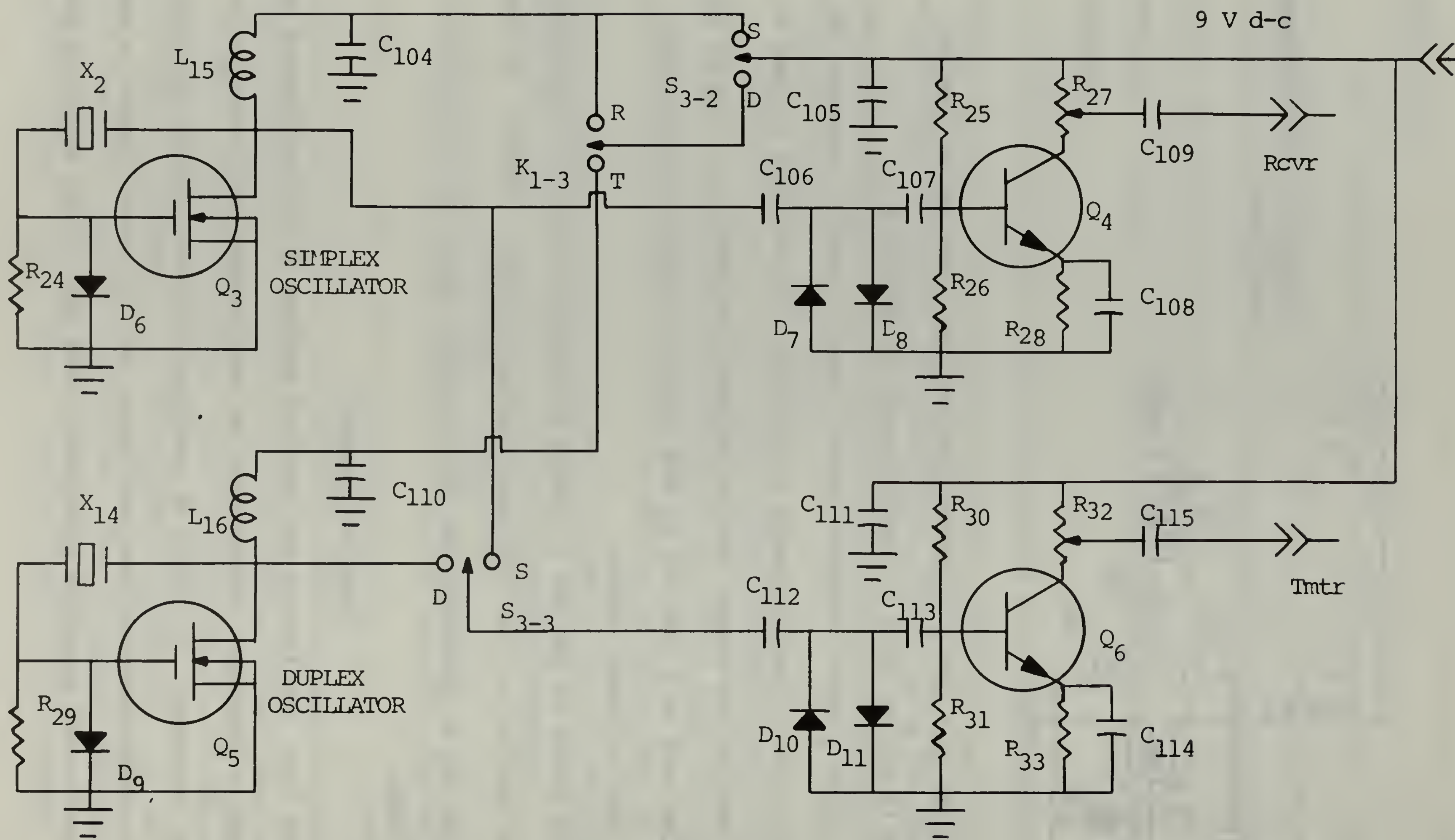


Figure 20. Local Oscillator.

at high frequency. It was felt that it would be easier and cheaper to use two oscillators and do most of the needed switching in d-c circuits. There is, however, still some high-frequency switching required.

Figure 20 shows the local oscillators with most of the required switching. The crystal switching is not shown here because it is quite simple and would only clutter up the schematic. A detailed description of the switching will be presented in the next section.

Both oscillator-buffer amplifier circuits are identical, so only one need be discussed. The FET, Q_5, in the duplex oscillator is biased by the gate biasing diode, D_9, and the gate-leak resistor, R_{29}, in the same manner as the BFO. In fact both circuits are the same except that the local oscillator circuits do not require the capacitive voltage divider across the crystals since all frequencies are above 2 MHz. The output of the duplex oscillator is coupled through the d-c blocking capacitors, C_{112} and C_{113}, to the base of the buffer amplifier. Diodes D_{10} and D_{11}, between C_{112} and C_{113}, serve to limit the input to the buffer amplifier to a fixed, constant level of approximately 1.2 V peak-to-peak. The gain of the amplifier is constant over the frequency range of interest so that the output level, once set by adjusting R_{32}, is constant for all channels. This is required in order that the mixer behave the same for all channels. The output is coupled through the d-c blocking capacitor, C_{115}, to the transmitter mixer. R_{30}, R_{31}, and R_{33} bias the amplifier for class A operation. C_{110}, C_{111}, and C_{114} are r-f by-pass capacitors and L_{16} is an r-f choke which keeps the r-f signal from getting into the d-c power supply.

SWITCHING

The transmitter switching involves d-c power distribution, channel selection, and such features as selection of the AM mode for transmission and reception on channel 1, automatic reduction of power to 60 W of carrier in AM, and reduction of power to 150 W PEP for ship-to-ship transmissions in band 1 (channels 1, 2, and 3). Either the AM mode or the SSB mode may be used for ship-to-ship transmission. Therefore the power must be reduced in these modes in band 1. The power does not have to be reduced in the ship-to-shore mode. Automatic AM on channel 1 is incorporated to minimize the amount of switching that the operator must perform since channel 1 is the calling and emergency channel at 2182 kHz, and since most receivers monitoring this frequency are presently equipped only for AM.

The channel selector selects the proper crystal for the local oscillator and the proper tuned circuits at the mixer output and at the grid and plate of the power amplifier. It is also electrically inter-connected to the mode switch to provide the automatic mode selection required.

To assure that no redundant switching is built into the transmitter, one can develop a truth table from which the switching-logic equations can be found. These equations can then be reduced by Boolean algebra to their minimum form. The reduced logic equations will then show the minimum switching required.

To develop a truth table one must identify the inputs, over which one has control, and the desired outputs. The outputs will differ depending upon the mode selected and the channel selected. However, there are many channels for which the output will be the same except for frequency.

Since it is the automatic features which make the switching problem interesting, while channel selection alone is trivial, one can simplify the truth table by lumping together all those channels for which the only difference is frequency. Thus the channel selector switch will reduce to three basic positions; i.e. a) channel 1, b) channels 2 and 3, and c) channels 4 through 12.

Define the mode positions as

AM for AM,

SS for ship-to-shore, and

SB for single sideband.

Further define the channel selector positions as

F1 for channel 1,

F2 for channels 2 and 3, and

F3 for channels 4 through 12.

The outputs will be identified mainly by their difference with respect to the basic single-sideband signal and by the power level. Define the basic single-sideband output to be

C0 for no carrier present, and

P0 for 250 W PEP output.

The other outputs in various combinations are

C1 for carrier level 16 dB below the PEP,

C2 for carrier level 6 dB below PEP,

P1 for 150 W PEP, and

P2 for 60 W carrier power.

Table I, the truth table, shows the various input conditions and the desired output conditions by using a 1 to indicate the presence of the specified condition and a 0 to indicate its absence. Note that the inputs refer to the positions selected by the channel selector and the mode selector on the front panel. The outputs are the results of this selection. As an example, line 2 in Table I indicates that the channel selector is set at channel 1 and the mode selector is in the ship-to-shore position. The output, however, is AM as indicated by the carrier level of -6 dB (C2).

INPUT						OUTPUT					
CHAN. SEL.			MODE SEL.			CARRIER LEVEL			POWER		
F1	F2	F3	AM	SS	SB	C0	C1	C2	P0	P1	P2
1	0	0	1	0	0	0	0	1	0	1	0
1	0	0	0	1	0	0	0	1	0	1	0
1	0	0	0	0	1	0	0	1	0	1	0
0	1	0	1	0	0	0	0	1	0	1	0
0	1	0	0	1	0	0	1	0	1	0	0
0	1	0	0	0	1	1	0	0	0	1	0
0	0	1	1	0	0	0	0	1	0	0	1
0	0	1	0	1	0	0	1	0	1	0	0
0	0	1	0	0	1	1	0	0	1	0	0

Table I. Truth table.

The desired characteristics of the output signal (relative carrier level and power) are accomplished independently within the transmitter, hence can be considered separately in writing the Boolean expressions.

From the truth table one can see that:

$\underline{C0} = (\underline{F2})(\underline{SB}) + (\underline{F3})(\underline{SB}) = (\underline{F2} + \underline{F3})(\underline{SB}),$

$\underline{C1} = (\underline{F2})(\underline{SS}) + (\underline{F3})(\underline{SS}) = (\underline{F2} + \underline{F3})(\underline{SS}),$

$\underline{C2} = (\underline{F1})(\underline{AM}) + (\underline{F1})(\underline{SS}) + (\underline{F1})(\underline{SB}) + (\underline{F2})(\underline{AM}) + (\underline{F3})(\underline{AM})$

$\quad = \underline{F1} + \underline{AM}^{*},$

$\underline{P0} = (\underline{F2})(\underline{SS}) + (\underline{F3})(\underline{SS}) + (\underline{F3})(\underline{SB})$

$\quad = (\underline{F2} + \underline{F3})(\underline{SS}) + (\underline{F3})(\underline{SB}),$

$\underline{P1} = (\underline{F1})(\underline{AM}) + (\underline{F1})(\underline{SS}) + (\underline{F1})(\underline{SB}) + (\underline{F2})(\underline{AM}) + (\underline{F2})(\underline{SB})$

$\quad = \underline{F1} + (\underline{F2})(\underline{AM} + \underline{SB}),$

$\underline{P2} = (\underline{F3})(\underline{AM}).$

These Boolean expressions for the desired outputs can now be used to check that the switching shown in subsequent diagrams is the minimum required to provide the desired external simplicity of operation.

CHANNEL SELECTION

Channel selection is accomplished by the channel selector switch on the front panel. This switch selects the desired crystal or crystals for insertion in the local oscillators. It also selects the proper coil and capacitor at the output of the final mixer, the proper transformer and capacitor at the grid of the power amplifier, and the proper capacitors and coil tap in the pi network.

* $\underline{F1} + \underline{F2} + \underline{F3} = 1$ since these are all the possible channels; and $\underline{AM} + \underline{SS} + \underline{SB} = 1$ since these are all the possible positions on the mode switch.

Figure 21 shows the switching of the tuned circuits at the output of the final mixer. As can be seen, it is rather a straightforward arrangement. One wafer of the channel selector switch is required for selecting the coil and another for selecting the capacitor. The cold end of each coil and tuning capacitor, C_{18} through C_{29}, are tied together at the top of C_{17}.

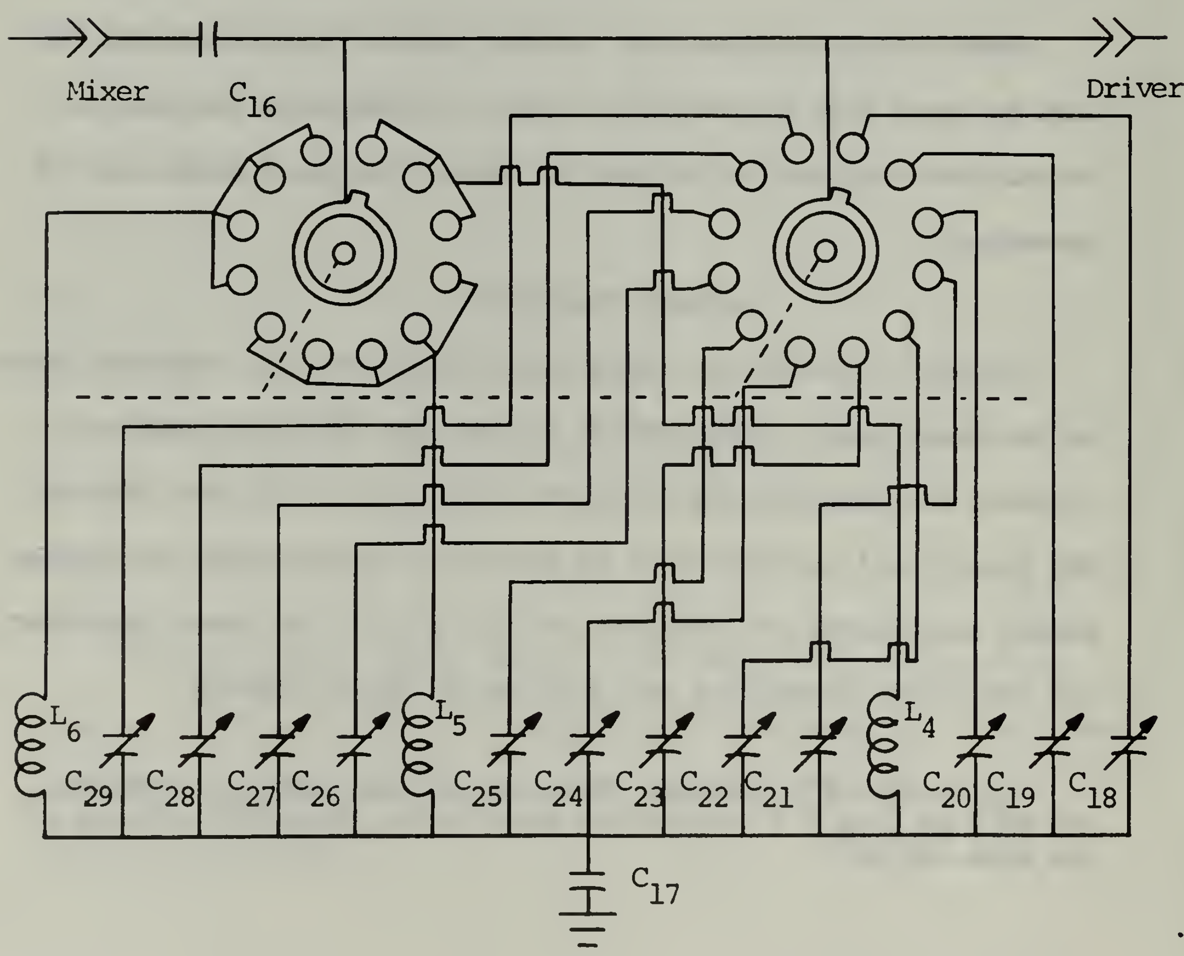


Figure 21. Channel selection at the mixer output.

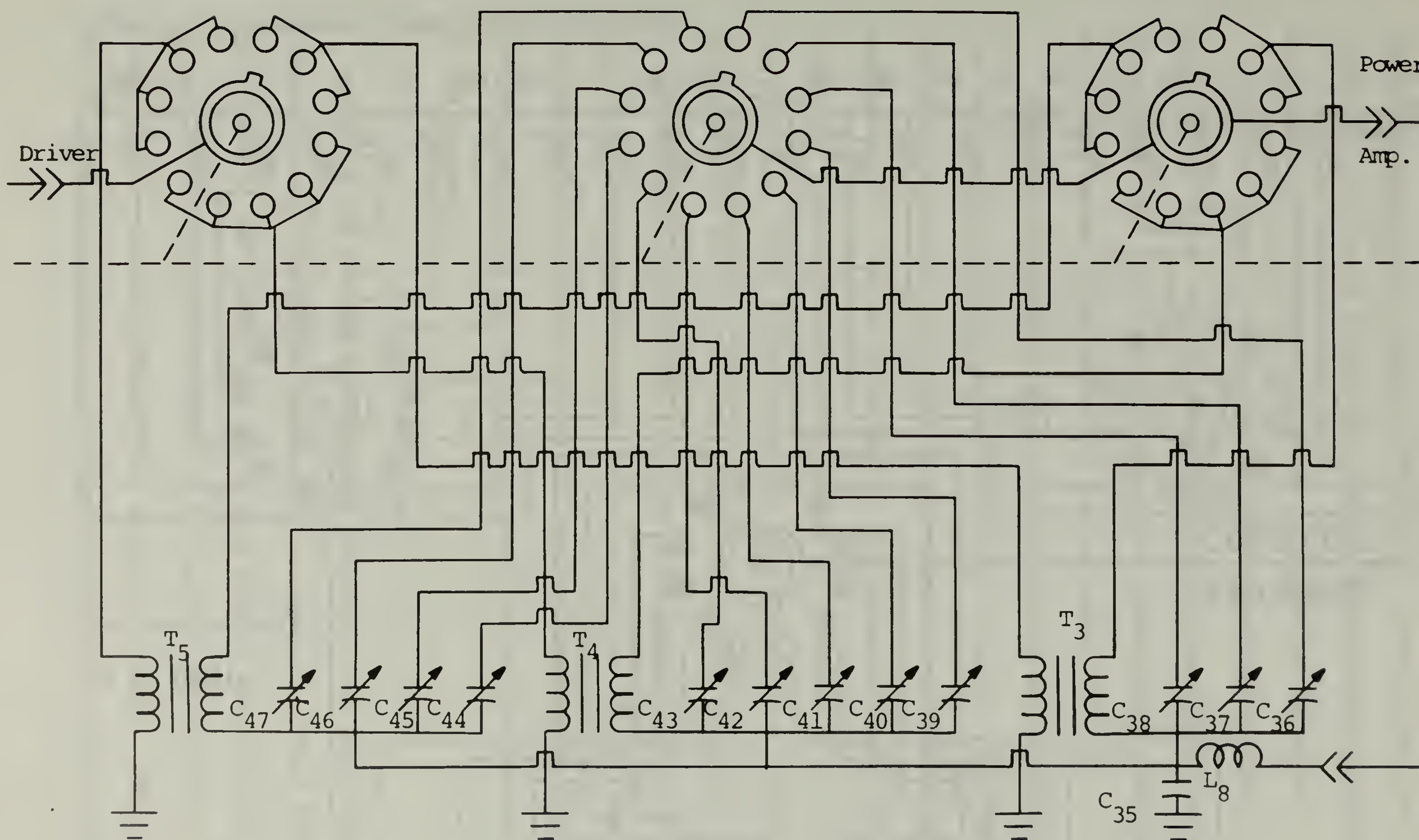


Figure 22. Tuned transformer selection at the grid of the power amplifier.

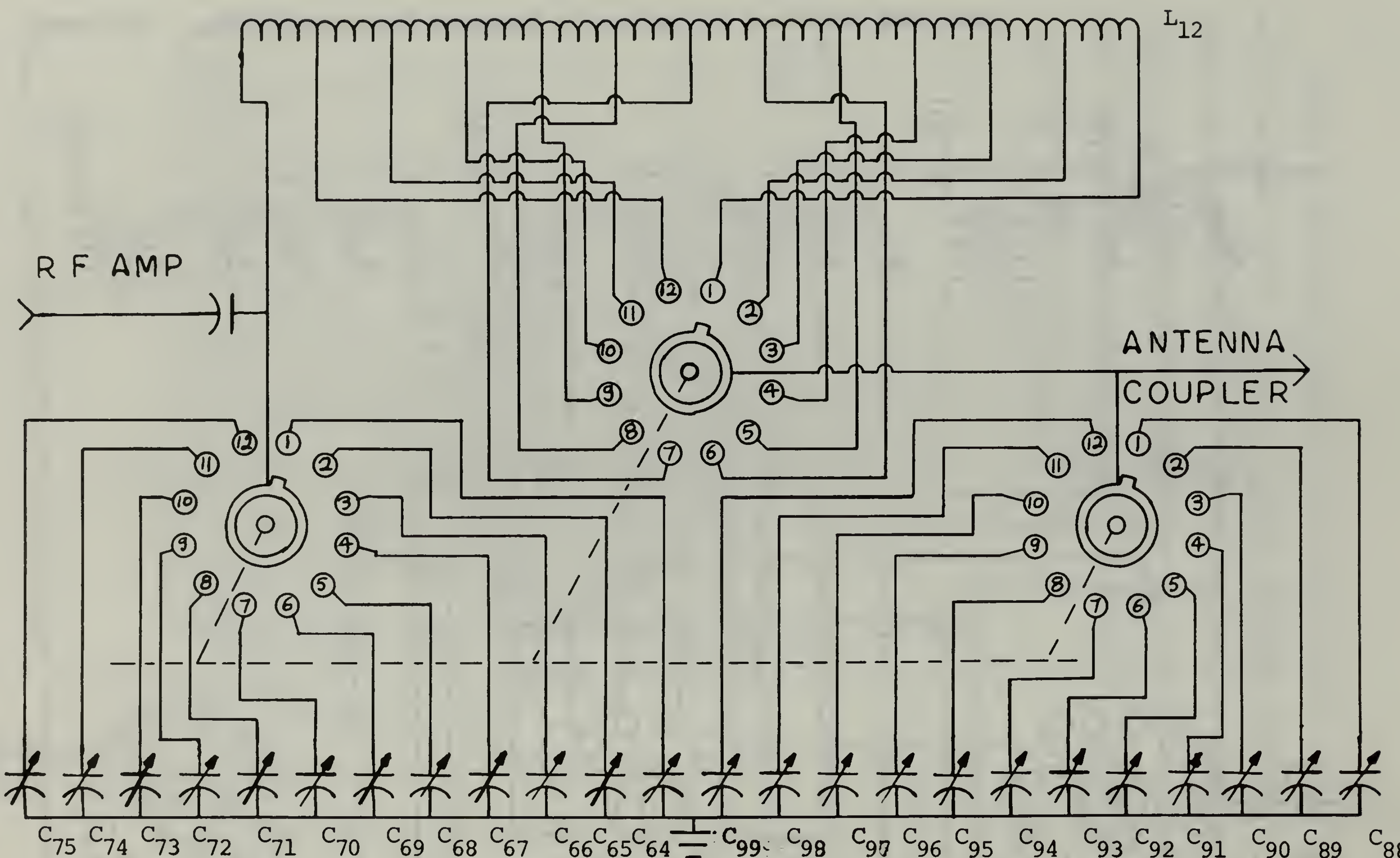


Figure 23. Pi-network tuned circuit selection.

The switching arrangement for selection of the transformer and tuning capacitor is shown in Figure 22. Three wafers are required here, one to select the primary of the desired transformer, one to select the secondary, and one to select the capacitor.

Finally Figure 23 shows the switching required in the pi network. Three wafers are also required here for selecting the input capacitance, the coil tap, and the output capacitance. The fixed capacitors at the input and output are not shown.

There are two ways in which the fixed capacitors can be inserted into the circuit. The first possibility is to simply place one fixed capacitor in parallel with each tuning capacitor and select both simultaneously. The second method is to have an array of fixed capacitors from which the desired capacitor or capacitors is selected. The advantage of the second method is that fewer fixed capacitors are required since two or more small capacitors may be selected in parallel to make a larger capacitance. However this requires two or more additional wafers on the channel selector switch. Since this switch already has a total of 24 wafers, 11 for the transmitter, 8 for the receiver, and 5 for the oscillators, the first of the possible methods is the one used.

AUTOMATIC AM

Most of the switching complexity resulting from the automatic-AM-on-channel-1 feature is related to the receiver. There are, however, two places in the transmitter where this feature must be taken into account. The switching arrangement for carrier reinsertion is one of these places. Figure 24 shows this switching network. Were it not for the fact that positive grounding of the carrier signal is required to prevent the carrier from leaking into the i-f amplifier in the SSB mode, only one channel selector wafer and one section of one wafer of the mode selector switch would be required. Since positive grounding is required, two channel selector wafers and all three sections of the mode selector wafer are required.

The second place where automatic AM is used is in the automatic power reduction feature discussed below.

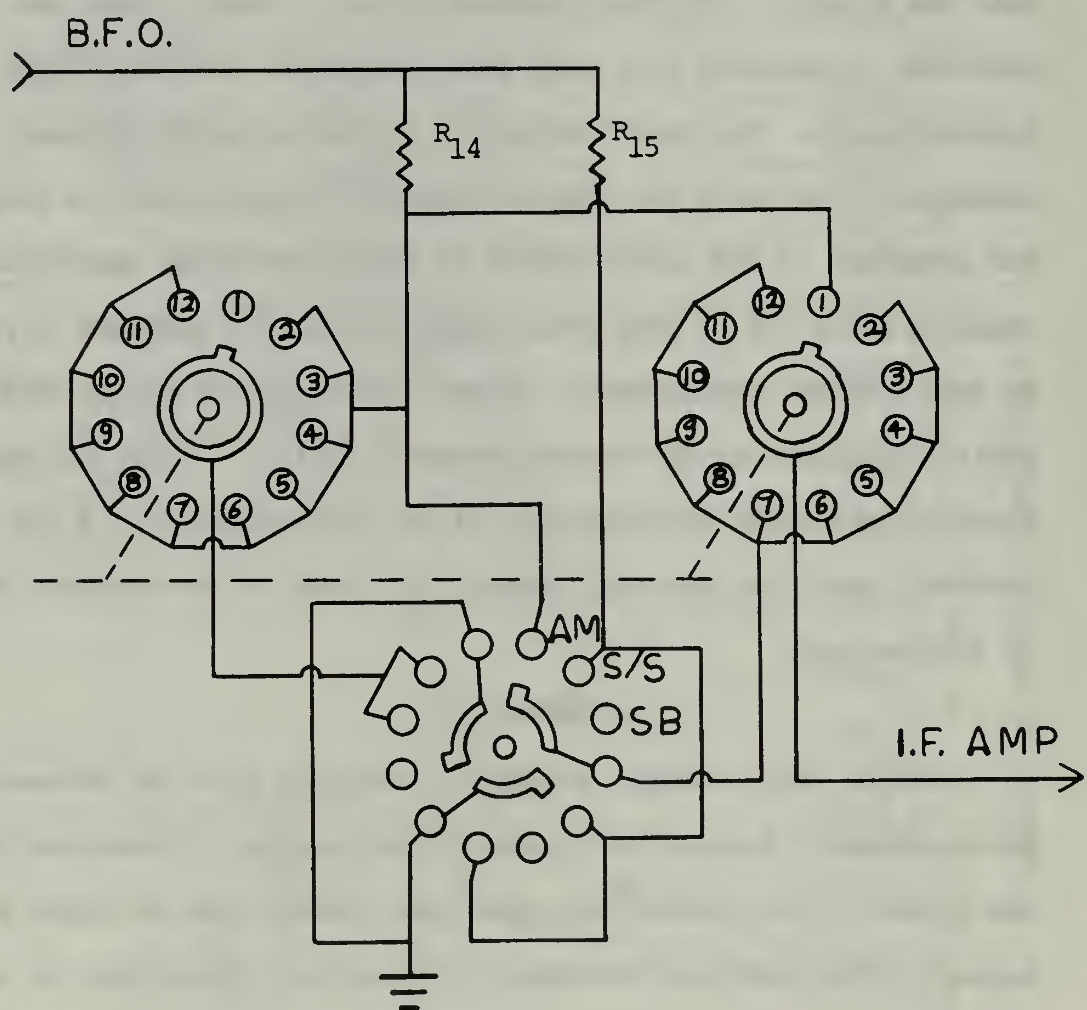


Figure 24. Switching network for carrier reinsertion.

AUTOMATIC POWER REDUCTION

Automatic power reduction is proposed to be accomplished by a switching network which switches the three threshold setting components into the ALC circuit as required by Table I. Since these threshold devices and the ALC circuit have not been established, the proposed switching network is presented in Figure 25 using the symbology of the Boolean equations. Thus P0 represents the threshold level for 250 W PEP output, P1 represents the threshold level for 150 W PEP output, and P2 represents the threshold level for 60 W carrier output.

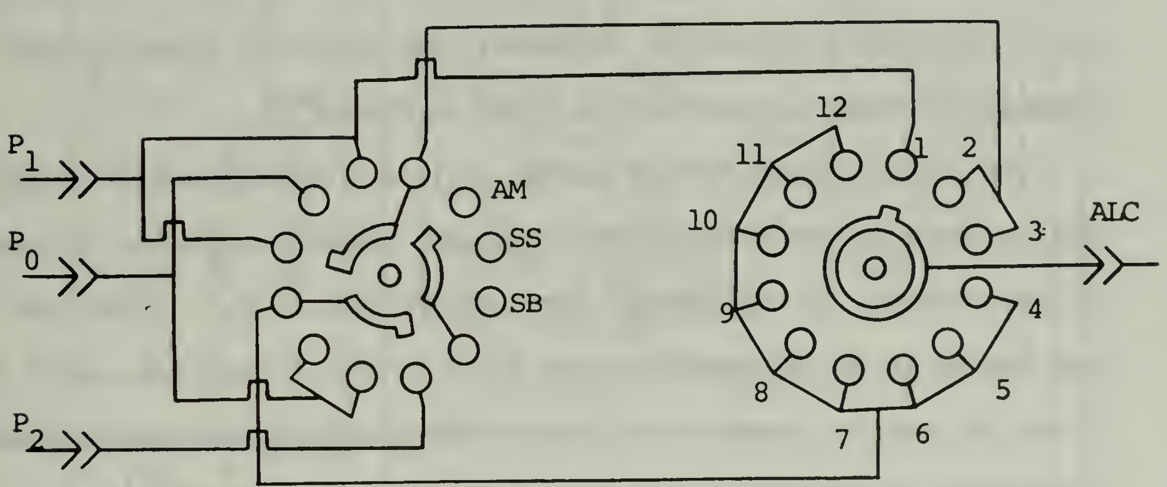


Figure 25. Switching network for automatic power reduction.

As can be seen in Figure 25, the threshold level for P0 is connected to the ALC line whenever the mode switch is in the ship-to-shore (SS) position or the single-sideband (SB) position and the channel selector set for any channel from 4 through 12. P0 is also obtained with the mode switch in the SS position and the channel selector in either channel 2 or 3.

The threshold level for P1 is obtained with the mode switch in either the AM or the SB position and the channel selector set for either channel 2 or 3. It is also obtained with the channel selector at channel 1 regardless of the position of the mode switch.

Finally the threshold level for P2 is obtained with the mode switch in the AM position and the channel selector set for any channel from 4 through 12.

D-C POWER DISTRIBUTION

The transceiver requires an external power supply which must have the following d-c voltages available; 850 V, 250 V, -50 V, and 12 V. Any other voltage levels required are derived from these four within the transceiver. Most of the front panel switches are concerned with the distribution of these d-c voltages. The switching network which accomplishes this distribution is shown in Figure 26.

The volume control/ON-OFF switch (S_4) is the basic switch which applies power to the transceiver. Only the crystal ovens may be turned on independently by the crystal ovens ON-OFF switch (S_5). Note that S_5 need not be in the ON position when S_4 is in the ON position. With S_4 in the ON position power is applied to the receiver, the crystal ovens (regardless of the position of S_5), and to the oscillators. 12 V is applied to the receiver circuit board through contact K_{1-1} of relay K_1 with no power applied to the relay coil.

A 9.1 V buss for the oscillators is developed by the Zener diode Z_2 in series with R_{33}. This regulated 9.1 V buss is provided to assure frequency stability and a constant level output from the oscillators. Since BFO injection to the receiver is not desired in the AM mode, the power to the BFO must be switched off whenever channel 1 or AM is selected.

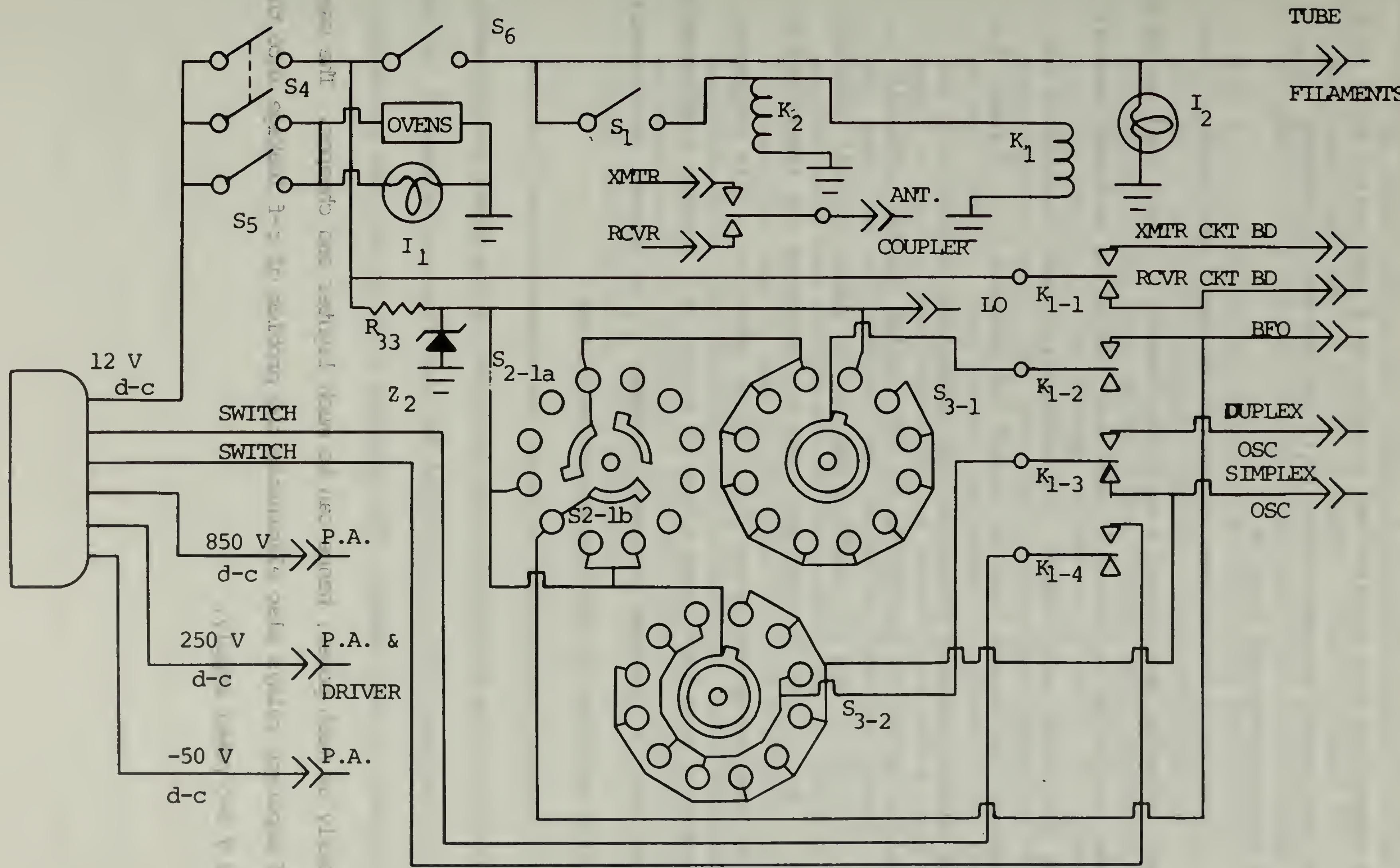


Figure 26. DC power distribution.

BFO injection is required for transmission. The switching arrangement to accomplish this requires two sections of a mode switch wafer (S_2), and one channel selector wafer (S_3), and one set of contacts (K_{1-2}) of the press-to-talk relay K_1.

Power at 9.1 V is also applied to the local oscillators in two ways; directly and through a switching network. Direct application of power is for the buffer amplifiers only, while the switched power is for the actual oscillators. The switching network consists of one wafer of the channel selector switch and contacts K_{1-3} of relay K_1. The channel selector wafer has been effectively reduced to a two position switch by connecting all the simplex channel terminals (1, 2, 4, 5, 6, 9, and 10) together and all the duplex channel terminals (3, 7, 8, 11, and 12) together. The two positions are identified as S for simplex operation and D for duplex operation. This switching network is also shown more simply in Figure 20. Another simplex-duplex wafer must be provided as shown in Figure 20, to connect the transmitter-buffer amplifier to the appropriate oscillator.

Relay K_2 is provided to switch the antenna coupler from the receiver to the transmitter when energized by the press-to-talk switch S_1. A separate relay is provided for this function because of the high r-f power it must handle. The contacts of K_1 do not have to handle nearly as much power, hence can be much lighter and cheaper. The use of separate relays also eliminates the problem of r-f leakage into the 12 V d-c power supply.

The three high-voltage lines are connected directly to the power amplifier and the driver as needed. These voltages are switched on by the press-to-talk switch only when transmitting. When S_1 is closed, relays K_1 and K_2 are energized causing their contacts to go to the transmit position. This completes the circuit of the two input switch lines through relay contacts K_{1-4}, allowing current to flow through another relay within the external power supply itself. The power supply relay then completes the high-voltage circuits. Thus another requirement is placed on the power supply in that it must have a pair of switch lines and a relay which will apply the high voltage when the switch lines are closed. Power supplies with this feature are currently available in industry.

Note that only S_4 need be activated to place the receiver in operation while three switches must be activated for transmission. Indicator lamps are provided to show when power is applied to the crystal ovens and to the tube filaments.

CONCLUSION

The output of the final mixer met all specifications as to carrier suppression, undesired sideband suppression, and intermodulation distortion. Difficulty was encountered in the driver stage in attempting to amplify the desired sideband while relatively attenuating the local oscillator signal and the sum of the i-f and the LO signals. The difficulty encountered was the instability of the driver, even though an attempt to neutralize the circuit was made.

The failure of the neutralization attempt may be due, in part, to the very high Q tuned circuits at both the input and the output of the stage. The unloaded Q's were on the order of 200. Very high impedances (output impedance of the mixer and input impedance of the power amplifier) formed the loads for these tuned circuits. Therefore the loaded Q's remained high. This makes the neutralization point very critical and difficult to find and maintain. The high Q's were desired to obtain suppression of the LO signal with only two parallel tuned circuits and the pi network.

There are two possible methods one might pursue to obtain the necessary selectivity, gain, and stability. The first of these is to use transistor amplifier stages prior to the driver with lower Q tuned circuits. Then the Q of the tuned circuits in the driver may also be lower. The big disadvantage of this method is that for each additional tuned circuit used, more switching wafers would be required.

The other method is to replace the mixer with another balanced modulator, eliminating the LO signal immediately. If this were done, the selectivity of subsequent tuned circuits would not have to be as great since the frequency component nearest the desired sideband is already eliminated. The disadvantage here is that the second balanced modulator would have to be capable of operating in a balanced condition over a wide range of frequencies.

If sufficient time were available, these alternate methods could be tested and evaluated, and the best one chosen for implementation in the transmitter.

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APPENDIX I

SPECIFICATIONS

POWER OUTPUT	250 W PEP SSB 60 W Carrier AM

Power automatically reduced to 150 W on 2-4 MHz ship-to-ship channels as per FCC requirements.

MODES	A3a - Carrier reduced 16 dB below PEP for ship-to-shore telephone
	A3h - Carrier reduced 6 dB below PEP for AM (automatic on 2182 kHz)
	A3j - Carrier suppressed a minimum of 40 dB below PEP for SSB
TRANSMISSION on USB	All modes
FREQUENCY RANGE	2-17 MHz
NUMBER of CHANNELS	12
Band 1	3 channels 2-4 MHz 2 channels simplex and 1 channel semi-duplex for ship-to-shore telephone
Band 2	5 channels 4-8 MHz 3 channels simplex and 2 channels semi-duplex
Band 3	4 channels 8-17 MHz 2 channels simplex and 2 channels semi-duplex
POWER SUPPLY	Separate unit
CIRCUITRY	
Transmitter	All low power stages shall make full use of state-of-the-art semiconductors including transistors and integrated circuits.
Receiver	Full complement of semiconductors.

	Dual FET front end with integrated circuit i-f and a-f amplifiers. Push-pull audio output for low power consumption. Speaker integral with disabling switch on hand set jack.
APPROXIMATE SIZE	Length - 18" Height - 8" Depth - 10"

APPENDIX II

PI NETWORK DESIGN

As previously mentioned in the section on the power amplifier, the pi network must perform two functions; (1) match the output impedance of the power amplifier tubes to the input impedance of an antenna coupler, and (2) suppress harmonics of the desired r-f signal. To accomplish this one can usually continuously vary all three of the basic elements of the pi network. In this project, however, the inductor, L_{12} of Figure 17, can be varied in discrete steps only since it is a tapped coil wound on a powdered iron toroidal core. This then requires that another parameter be varied continuously. The parameter chosen was the input quality factor (Q_i) of the pi network.

If the frequencies of operation were known in advance for a particular transmitter, the required values of L_{12} could be calculated with a fixed or constant Q_i. This practice is not feasible, however, for commercial production. In a radio designed for mass production which must cover a wide band of frequencies, a limited number of values for L_{12} must be chosen and each value must be capable of covering a band of frequencies. To do this Q_i must be allowed to vary. Limits must be placed on the variation of Q_i since too high a value results in low efficiency and too low a value results in poor harmonic suppression. In the design of this pi network, Q_i was allowed to vary from a high of 15 at the low end of the particular frequency band, to a low of 12 at the high end of the frequency band.

There are other restrictions affecting the pi network design which are more physical in nature and which may be the overriding limitation. One of these restrictions is the capacitive range of the input tuning capacitor, which in turn is restricted by the physical size limitations and the peak voltage which it must be capable of withstanding. The peak voltage which the input capacitor must withstand is twice the plate supply voltage, in this case 1700 V. This suggests that the input tuning capacitor be an air variable. The two air variable capacitors chosen to fit these limitations have a capacitive range of from 8 to 140 pF and 38 to 202 pF. Having two types of capacitors available provides greater design flexibility.

The tuning range of the input capacitor proved to be the most critical restriction over most of the frequency range of the transmitter. Thus a trial-and-error process of design had to be used. To facilitate this process, Table II, a table of the reactances of the three elements with Q_i being the variable parameter, was developed. The steps in the development of this table are given below.

1) Select the desired value of Q_i.

2) Knowing the desired input parallel resistance, R_{pi} (900 ohms in this case), calculate the susceptance of the input capacitor; $B_{ci} = Q_i/R_{pi}$.

3) Calculate the values of the equivalent series circuit; $R_{si} = R_{pi}/(Q_i^2 + 1)$ and $X'_{ci} = Q_i R_{si}$.

4) For an impedance match one wants the resistance of the equivalent series circuit of the output capacitor and resistive load (R_{po}), to be equal to the equivalent series resistance at the input. From this, one can calculate the quality factor of the output capacitor-load combination (Q_o); $Q_o = R_{po}/R_{si} - 1$.

Q_i	B_{ci}	R_{si}	X'_{ci}	Q_o	B_{co}	X'_{co}	X_L
15.0	0.0167	3.98	59.7	3.40	0.0697	13.5	73.2
14.9	0.0166	4.04	60.1	3.38	0.0676	13.6	73.7
14.8	0.0164	4.09	60.5	3.35	0.0670	13.7	74.2
14.7	0.0163	4.15	61.0	3.33	0.0666	13.8	74.8
14.6	0.0162	4.20	61.4	3.30	0.0660	13.9	75.3
14.5	0.0161	4.26	61.8	3.28	0.0656	14.0	75.8
14.4	0.0160	4.33	62.4	3.25	0.0650	14.05	76.45
14.3	0.0159	4.39	62.7	3.22	0.0644	14.1	76.8
14.2	0.0158	4.44	63.0	3.21	0.0642	14.2	77.2
14.1	0.0157	4.50	63.5	3.18	0.0636	14.3	77.8
14.0	0.0156	4.57	64.0	3.15	0.0630	14.4	78.4
13.9	0.0154	4.64	64.4	3.13	0.0626	14.5	78.9
13.8	0.0153	4.70	64.9	3.10	0.0620	14.6	79.5
13.7	0.0152	4.76	65.2	3.08	0.0616	14.7	79.9
13.6	0.0151	4.84	65.7	3.05	0.0610	14.75	80.45
13.5	0.0150	4.91	66.4	3.04	0.0608	14.9	81.3
13.4	0.0149	4.99	66.8	3.00	0.0600	14.95	81.75
13.3	0.0148	5.05	67.1	2.98	0.0596	15.05	82.15
13.2	0.0147	5.11	67.5	2.96	0.0592	15.1	82.6
13.1	0.0146	5.20	68.0	2.94	0.0588	15.3	83.3
13.0	0.0144	5.30	68.9	2.91	0.0582	15.4	84.3
12.9	0.0143	5.39	69.5	2.88	0.0576	15.5	85.0
12.8	0.0142	5.45	69.8	2.86	0.0572	15.6	85.4
12.7	0.0141	5.55	70.5	2.85	0.0570	15.8	86.3
12.6	0.0140	5.63	70.9	2.81	0.0562	15.8	86.7
12.5	0.0139	5.74	71.6	2.78	0.0556	15.9	87.5

Q_i	B_{ci}	R_{si}	X'_{ci}	Q_o	B_{co}	X'_{co}	X_L
12.4	0.0138	5.80	72.0	2.76	0.0552	16.0	88.0
12.3	0.0137	5.92	72.9	2.73	0.0546	16.2	89.1
12.2	0.0136	6.00	73.2	2.72	0.0544	16.4	89.6
12.1	0.0134	6.10	73.8	2.68	0.0536	16.4	90.2
12.0	0.0133	6.20	74.4	2.66	0.0532	16.5	90.9

Table II. Table of pi network component reactances.

5) Find the susceptance of the output capacitor; $B_{co} = Q_o/R_{po}$.

6) Find the reactance of the equivalent series capacitor at the output; $X'_{co} = R_{si}Q_o$.

7) Finally, find the required inductive reactance for series resonance of the equivalent of the pi network; $X_L = X'_{ci} + X'_{co}$.

Now that Table II has been compiled one can proceed with the trial-and-error design of the pi network. Begin at the low end of the frequency range (2 MHz) and calculate the values of the input capacitance (C_i), the inductance (L), and the output capacitance (C_o) corresponding to the tabled values of B_{ci}, X_L, and B_{co} for the maximum Q_i. The values found for C_i and C_o will be the maximum of their tuning range; call them C_{ih} and C_{oh}. The next step is to select a lower value of Q_i, find the corresponding value of X_L and, using the previously calculated value of L, calculate the frequency. Now using this new frequency and the B_{ci} corresponding to the lower Q_i, find C_i. This new value of C_i should be the lower limit of the input capacitor, call it C_{il}. If the difference of C_{ih} and C_{il} is greater than the maximum tuning range of the input capacitor, a higher value of Q_i must be chosen and C_{il} recomputed. If the difference is less than the tuning range of the input capacitor, a lower value of Q_i may be chosen.

This process is repeated until the difference between C_{ih} and C_{il} corresponds to the tuning range of C_i or until the lower limit of Q_i is reached. When this equilibrium is reached C_{ol} may be calculated using the same high frequency, and the limits of C_o checked. In this problem C_i was much more restrictive. The entire process is now repeated for the next band of frequencies starting with the maximum Q_i and the last value of frequency calculated above. Table III shows the result of this process.

To illustrate the procedure, the first entry of Table III will be calculated. From Table II, the values of B_{ci}, B_{co}, and X_L for $Q_i = 15$ are, 0.0167 mho, 0.0697 mho, and 73.2 ohms, respectively. The corresponding values of C_{ih}, C_{oh}, and L are 1330 pF, 5550 pF, and 5.84 uH for $\omega = 12.56$ Mrad. Choosing now a low value for Q_i of 14.0, from Table II the values of B_{ci}, B_{co}, and X_L are 0.0156 mho, 0.0630 mho, and 78.4 ohms, respectively. Since L remains fixed over this band use $L = 5.84$ uH and $X_L = 78.4$ ohms to find the high frequency of 13.42 Mrad. Now using $\omega = 13.42$ Mrad and $B_{ci} = 0.0156$ mho, $C_{il} = 1160$ pF. The difference between C_{ih} and C_{il} is 170 pF while the tuning ranges of the two air variable capacitors available are 132 pF and 164 pF. Thus the required tuning range is beyond the capability of either variable capacitor. Choosing now a higher value for Q_i, say $Q_i = 14.4$, then $B_{ci} = 0.0160$ mho, $B_{co} = 0.0650$ mho, and $X_L = 76.45$ ohms. Now $\omega = 13.1$ Mrad and $C_{il} = 1220$ pF so that the required tuning range is now 110 pF, well within the capability of either air variable. Another factor which must be considered in choosing which of the two air variable capacitors to use is the capacitance of available fixed capacitors.

To compute the next entry of Table III, start with a frequency of 13.1 Mrad and $Q_i = 15$ and repeat the above procedure. The entire process is then repeated until the maximum frequency of the transceiver is attained.

Table III lists the frequency range for each band in Mrad, the range of Q_i for each band, the range of the input and output capacitances required and the fixed portion of those capacitances, and the inductance for each band. Note that in the pi network it was not possible to use only three bands as was the case elsewhere in the transmitter.

This design procedure neglects losses in the components of the pi network. This is a reasonable assumption since very high Q components should be used in the network to keep power loss and heating in the network to a minimum.

ω (Mrad)	Q_i	C_i (pF)	C_{if} (pF)	L (uH)	C_o (pF)	C_{of} (pF)
12.6-13.1	15-14.4	1330-1220	1200	5.84	5550-4960	4700
13.1-13.8	15-14.3	1275-1155	1100	5.59	5320-4680	4300
13.8-14.3	15-14.4	1215-1120	1100	5.33	5070-4550	4300
14.3-15.1	15-14.2	1170-1050	1000	5.12	4880-4250	3900
15.1-15.7	15-14.4	1105-1020	1000	4.85	4610-4140	3900
15.7-16.6	15-14.2	1060- 950	900	4.66	4450-3880	3600
16.6-17.3	15-14.3	1010- 920	900	4.42	4210-3720	3600
17.3-18.3	15-14.0	970- 853	800	4.24	4040-3440	3300
18.3-19.3	15-14.2	915- 818	800	4.00	3810-3320	3000
19.3-20.7	15-14.0	865- 750	700	3.79	3620-3040	2700
20.7-21.8	15-14.2	807- 725	700	3.54	3370-2940	2700
21.8-23.5	15-13.9	766- 650	600	3.36	3200-2660	2400
23.5-25.1	15-14.0	711- 620	600	3.12	2970-2510	2400
25.1-27.5	15-13.6	665- 550	500	2.92	2780-2220	2000
27.5-29.6	15-13.9	608- 520	500	2.66	2540-2110	2000
29.6-33.1	15-13.4	564- 450	400	2.47	2360-1810	1600
33.1-36.2	15-13.7	505- 420	400	2.21	2110-1700	1500
36.2-41.2	15-13.2	462- 354	300	2.02	1930-1430	1200
41.2-47.7	15-12.9	405- 300	250	1.78	1690-1200	1000
47.7-57.5	15-12.4	350- 240	220	1.53	1460- 960	750
57.5-71.5	15-12.0	291- 186	150	1.27	1210- 744	500
71.5-89.1	15-12.0	234- 149	100	1.02	975- 596	300
89.1-110	15-12.0	188- 121	0	0.822	783- 484	0

Table III. Table of pi network component values.

APPENDIX III

PARTS LIST

Capacitors

All capacitors are 20% tolerance, disc ceramic unless specified otherwise.

C_1	0.02	uF			
C_2	1.0	uF	electrolytic	25	V
C_3	2.2	uF	electrolytic	25	V
$C_{4,13-16,31,32,104-114}$	0.1	uF		75	V
C_5	0.2	uF			
C_{6-8}	22	uF	electrolytic	25	V
C_9	200	pF	silver mica		
C_{10}	150	pF	silver mica		
C_{11}	7-60	pF	mica variable, Arco	404	
C_{12}	0.01	uF		75	V
$C_{18-29,102}$	14-150	pF	mica variable, Arco	424	
$C_{33,34}$	0.1	uF		1	kV
C_{35}	910	pF			
C_{48}	20	pF	variable		
C_{49}	0.01	uF		1	kV
C_{50}	0.001	uF		1.5	kV
C_{51}	0.001	uF		2.5	kV
C_{100}	51	pF			
C_{101}	15	pF	silver mica		
C_{103}	100	pF	silver mica		

C_{52-99}	values dependent upon channel frequencies. See Table III.			
$C_{17,30,36-47}$	values not determined			
Diodes				
D_{1-4}	1N34A			
Z_1	1N827			
$D_{5,6,9}$	1N4152			
$D_{7,8,10,11}$	1N929			
Z_2	1N3019A			
Filter				
F_1	1650 kHz crystal filter, Filtech Corporation Model S204165-L			
Inductors				
$L_{1-3,7,8,13,15,16}$	10	mH	molded RFC	100 mA
L_4	39.5	uH	toroid: 96 turns No. 32, enameled copper wire	
L_5	10.8	uH	toroid: 45 turns No. 28, enameled copper wire	
L_6	2.35	uH	toroid: 18 turns No. 28, enameled copper wire	
$L_{9,11}$			10 turns No. 20 enameled copper wire wound around a 47 ohm, 1 W resistor	
L_{10}	50	mH	RFC	100 mA
L_{12}			refer to Table III	
L_{14}	120-330	mH	Miller 4208	

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ABSTRACT

Design of the transmitter portion of a solid state, state of the art, single sideband, 2 - 17 MHz transceiver is presented. A short discussion of the theory of single sideband and a comparison of single sideband systems with amplitude modulated systems are also presented. The unique requirements of commercial, marine communications are considered and the method of their fulfillment in this transceiver is discussed. Circuitry common to both the receiver and the transmitter is presented in detail.

14 KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT

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